Proceedings of the I·R·E

Journal of Communications and Electronic Engineering

January, 1950

Volume 38 Number 1

National Bureau of Standards

ELECTRON MICROSCOPY OF MAGNETIC FIELDS

n electron-optical shadow method, magnetic field configurations are
sible, and their field intensities become computable.

PROCEEDINGS OF THE I.R.E.

TABLE OF CONTENTS, INDICATED BY BLACK-AND-WHITE MARGIN, FOLLOWS PAGE 32A

The Institute of Radio Engineers

Changes of address (with advance notice of fifteen days) and communications regarding subscriptions and payments should be mailed to the Secretary of the Institute, at 450 Ahnaip St., Menasha, Wisconsin, or 1 East 79 Street, New York 21, N. Y.

PROCEEDINGS OF THE I.R.E.

Published Monthly by

The Institute of Radio Engineers, Inc.

VOLUME 38 — *January, 1950* — NUMBER 1

PROCEEDINGS OF THE I.R.E.

INSTITUTE NEWS AND RADIO NOTES SECTION

Responsibility for the contents of papers published in the PROCEEDINGS OF THE I.R.E. rests upon the authors. Statements made in papers are not binding on the Institute or its members.

Raymond F. Guy

PRESIDENT, 1950

Raymond F. Guy was a radio amateur from 1911 until 1916 when he joined the professional ranks as a ship's radio officer and radio inspector. He was employed at intervals by the Marconi Wireless Telegraph Co., the Independent Wireless Telegraph Co., and the Shipowners Radio Service. During the concluding engagements of World War I, he served in France in the Signal Corps of the Regular Army of the United States. Upon being discharged he entered Pratt Institute, graduating with the electrical engineering degree in 1921.

The New Year marks Mr. Guy's thirtieth consecutive year as a broadcast engineer. He became a member of the staff of WJZ when, in 1921, it began operations as the world's second licensed broadcasting station. In 1924 he became a member of the engineering staff of the RCA Research Laboratories, supervising engineering, development, and construction of standard and short-wave broadcasting apparatus, stations, and systems, and participating in RCA's earliest television development.

In 1929 Mr. Guy transferred to the National Broadcasting Company to direct its frequency allocations engineering, and the planning, design, and construction of all NBC transmitting facilities. His activities have included all technical and allied phases of the industry during the entire growth period of AM, FM, short-wave, and television broadcasting. During World War II, Mr. Guy participated in projects of the OSS, the CIAA, and the OWI. Since then he has taken part in International Conferences on radio in Havanna, Mexico City, Montreal, and Washington.

Mr. Guy, who is Manager of Radio and Allocations Engineering for the National Broadcasting Company, became an IRE Associate Member in 1925; a Member in 1931; and a Fellow in 1939. In 1943 he was elected to the Board of Directors and served through 1948, including one term as treasurer. Mr. Guy has been Chairman of the Standards, Public Relations, Founders, Transmitters, Membership, and Office Practices Committees, and Vice-Chairman of the Building Fund and Executive Committees. He has also served on Several Panels and Committees of the Radio Technical Planning Board and has represented the Institute in the activities of the American Standards Association.

His other affiliations include membership in the Society of Professional Engineers and Radio Executives Club; charter membership of the Radio Pioneers; and life membership of the Veterans Wireless Operator's Association. He is a Fellow of the Radio Club of America.

In the December, 1948, issue of the PROCEEDINGS, page 1451, there appeared a guest editorial by Mr. C. W. Carnahan entitled "Are You Satisfied?" In this, the relative advantages of governmental employment for engineers, as compared to private employment, were stressed.

Proposed guest editorials descriptive of the advantages of industrial employment for engineers, on the other hand, have been independently received from Messrs. B. B. Bauer (of Shure Brothers, Inc., Chicago, Ill.) and E. W. Butler (of Federal Telephone & Radio Corporation, Clifton, N. J.).

In pursuance of the policy of the Institute to present relevant aspects of suitable controversial questions, and in view of the intrinsic interest of these guest editorials to the membership, they are here presented.—*The Editor*

The Engineer in Private Enterprise

C. W. Carnahan, writing in the PROCEEDINGS OF THE I.R.E., pointed to certain factors which he felt should induce engineers to enter Government service. As one who has spent his entire business life in private industry, I believe we should also review the advantages to the engineer of entering private industry.

Obviously, the reasons for any individual's choice will lie largely in his personal views on what will give him satisfaction from his life's work. To get such satisfaction he must have a feeling that what he is doing is really justifying his position in the scheme of things. It is too casually accepted by many of us that the economic system within which we live has provided the highest material standard of living the world has seen. We are prone to overlook the fact that fundamental opportunities for individual growth and progress are at the roots of this great system.

What does this mean to the engineer?

Among the many benefits offered by private enterprise, he should consider that in it he became part of a business team. Such teams by the tens of thousands have conceived, engineered, produced, marketed, and financed a variety of products from television to mousetraps. These products have been within the pocketbook range of millions, and have created employment for millions at the world's highest real wages.

What position can the engineer play in the business team of engineering, marketing, production, and finance? The answer is that he can play almost any part. Not every graduate engineer wishes to do research, development, or design. The free enterprise system, private industry, gives him the chance to apply his engineering training to sales, production, or finances, should he prefer. Naturally he will want to study further if he decides to follow other than strict engineering, but the needs and opportunities are there.

The technological growth of our world gives an advantage to the man who has engineering training. He can play the "marketing" position, where he will help determine what product the customer wants, what features he prefers in it, what prices he will pay, and the quantity that will be bought. He can play the "engineering" position, where he will work with his associates in marketing and production to develop a product which will offer the prospective customer the best "package" of features and price. He can play the "production" position, working with marketing, engineering, and finance to get cost down and quality up. Or in the "financial" position he can, with the advantage of engineering knowledge, work with his other team mates in providing the capital which will cover the entire program from product conception to the customer's door.

Everyone on a business team is working to give a better product and a better value than competitors offer. This joint effort means financial success and personal growth for the alert, energetic, and enterprising. For the customer, it means better products and better values.

Engineers who enjoy teamwork and exciting competition can hardly ask for more than to be part of such a team in our free enterprise system.—*E. W. Butler*

Because I expected that others would challenge the editorial "Are You Satisfied?" by C. W. Carnahan, I have waited until now to voice my opinion.

To begin with, I do not doubt that in many organizations the "battle with the sales department, whose decisions and salaries invariably outrank their own" can be highly irritating to the engineer. There are other human relationships in private industry that are unpleasant; but I am inclined to believe that even engineers and scientists in the government service do not work under utopian conditions, as witness the attacks upon Dr. E. U. Condon and others.

My basic difference with Mr. Carnahan is that I cannot share his attitude toward competition and profits and our system of free market enterprise. We must remember that in production for war, we utilized the mechanisms of production that had been developed to serve an economy in which competition had been the anvil upon which mass production know-how was forged. Something certainly must be said for an industrial system which started from scratch and provided a flow of munitions and supplies which became the basis of our victory.

While the war effort was " production at any cost," we, as engineers, should be the first to recognize that such a basis encourages waste and inefficiency. If we are to produce more goods for the most people, we *must* be concerned about the cost of those goods to the consumer, and find the need for cost reduction a valid engineering challenge. Today, that challenge to production is more than a mere technical need. It is even more than a competitive requirement. It is the foundation of the very existence of our way of life. The way to a stable economy must not be lower wages per hour, but lower costs per unit of production, and it should appeal to engineers that this peacetime challenge can be as vital and stimulating as production for victory in war.

I hope the time will never again come when competition will be fought out on the basis of the ability of industry to hire men for less or to drive them harder. I, for one, prefer the competitive challenge for the fulfillment of human wants and for lower costs to be a matter of concern to engineers. To the degree which he accepts and meets that challenge, the engineer will rise to leadership in the executive councils of his company.

It is to be hoped that engineers will continue to meet the important needs of our economy. It is to be hoped that their unconscious desire to be sheltered from competitive drive will not become their conscious aim.

Capable engineers are needed in both government jobs and in industry. Many capable engineers are in our government service, and they are serving our country as effectively as many in private industry. It is my feeling, however, that men who enter government service merely for the reasons presented by Mr. Carnahan are not likely to be any more effective in government service than they were in industry.—*B. B. Bauer*

Service Engineering for Television*

E. EUGENE ECKLUND†

Summary—**Shortcomings in service engineering cost the consumer tens of thousands of dollars per year. Although the Armed Forces did much to make the electronics engineer service-conscious, there is considerable evidence that these experiences are not being fully utilized in the television industry. A proper approach to this problem can be of immeasurable value to the television manufacturer.**

During recent years, there has been an increase in demand by the service technician for added attention to servicing problems during design and development stages. This demand was probably present for a great number of years, but due to the increased recognition of the importance of servicing, it has become more apparent at the present time. With the arrival of television, its added complications, and the customer demands for perfection, this problem has, in many cases, become very serious. The electronics industry, on the whole, is lacking considerably in a practical approach to this problem. It is conceded that manufacturers are more service-conscious than ever before, primarily due to the lessons learned during the recent war years, but there is still much to be desired.

In looking back at the progress of the electronics servicing business, it is evident that this phase of the industry has pulled itself up by its own bootstraps. It is time that the electronics industry gave it a push to help it along. A review of portions of the past may be helpful in establishing a course for the future.

Prior to the war, the manufacturing of test equipment and aids to servicemen was very small, and of a minor nature considering the over-all electronics field. At the outset of the war, there were only a relatively few types of test equipment being manufactured, and these were more simple in nature. Precision equipment for servicemen was virtually unknown, partly, of course, because of the lack of need for such equipment. Although customer complaints were given consideration, the various complaints of the serviceman were often either ignored, or conceded and subsequently neglected.

At the outset of the war, the average manufacturer was definitely not service-conscious. Specialized electronic equipment, particularly radar, was being developed at a rapid rate to fill the urgent requirements of operational necessity.

Government and industry alike were faced with a very serious problem of developing equipment of a very new nature to operate on frequencies heretofore considered only experimental, and there was a decided lack of qualified technical personnel. The lack of adequate test equipment added considerably to the seriousness of the situation. Only very simple field test equipment was available, most of which was not particularly accurate. Those equipments more accurate in nature were laboratory-type equipment which was always considered very delicate. A hurried development program was started, but was considerably behind the development of the basic equipment with which it was intended to be used. Design engineers admitted that equipment maintenance and servicing would be a difficult problem, but usually had no idea of how properly to test the equipment until laboratory designs of the basic equipment were completed. As a result, much of the test equipment design work was not pushed until the design of the basic equipment was "frozen."

In order to help overcome this situation, the Radiation Laboratory of the Massachusetts Institute of Technology instituted a program which was perfectly logical under the circumstances but nevertheless novel in its approach. This laboratory had one group which was devoted entirely to design and development of test equipment to be used with various radar equipments. Realizing the seriousness of the Armed Forces' maintenance problems, and of the difficulty of educating the design engineer in the need for better approaches, the Radiation Laboratory set up a liaison engineer to co-ordinate the work of the receiver group and the test equipment group. Fortunately, the Radiation Laboratory picked a very capable man of high intelligence and sound judgment. As later developments show, this was perhaps one of the major factors toward the eventual development of a sound and logical test equipment program. This arrangement immediately assisted the Services in educating the radar design engineer in the service problems and their approach to them. Another important step was taken when the Navy Bureau of Aeronautics, the Radiation Laboratory, and the Army Air Forces approached the Joint Radio Board of the Joint Aircraft Committee regarding standardization of test equipment programs. Since the Joint Radio Board was charged with the responsibility of co-ordinating aircraft radio matters between the Services, quick approval was granted and an *ad hoc* committee eventually consisted of representatives of the Radiation Laboratory, the Radio Research Laboratory of Harvard University, Navy Bureau of Aeronautics, Navy Bureau of Ships (Aircraft Electronic Section), Naval Research Laboratory, Co-ordinated Research Group, Aircraft Radio Laboratory of Wright Field, Army Air Forces, Royal Canadian Air Force, and Royal Air Force. Through this committee, the design and development of test equipment were controlled by assigning certain types of problems to one of the branches of the service as its prime responsibility. In addition, duplication of types of test equipment was virtually eliminated through agreement as to the equipment necessary to adequately service and maintain each of the various types of basic electronic gear. Also, a basic specification for test equipment was drawn up to cover the design and manufacture of all such products.

In co-ordinating the type of test equipment to be used, a specification was approved which stipulated the various voltages which should be provided in electronic equipment in the form of test points. These specifications assured that pertinent electronic information would be made readily available to the serviceman, allowing him to make a thorough check in a much shorter time. In addition, the limits of these test voltages were set forth, thus assisting the design engineer and assuring that standard test equipment could be used. Prior to this time, many engineers considered the same features to be important, but brought the test points out in various ways which they thought best. This often resulted in the necessity of designing special test equipment suitable for only the one basic set.

As the war developed, the American engineer provided many fine developments in test equipment which materially assisted the service technician. As time went on, industry became more and more aware of the needs for these developments and responded accordingly.

To depart from the American scene for a few moments, it is generally conceded that in most phases of electronic warfare, the enemy fell behind. In many cases, operational aircraft which had been shot down were investigated and found to have had all radio equipment removed even though it was normally of a standard type well-known to the Allies. Subsequent technical intelligence showed that, on the whole, the enemy, particularly in Europe, was equipped with very little test equipment. Such test equipment as was available, was primarily of the type used by the radio serviceman in the United States prior to the war. Although there was no definite basis, it is quite possible that the lack of consideration for the maintenance problem seriously affected the enemies' electronics picture.

The effect of electronic test equipment developments during the war has been readily apparent since that time. The cathode-ray oscillograph, which was a laboratory piece of equipment before the war and virtually unknown to the serviceman became a piece of standard operating equipment. With the advent of television, precision signal generators of a laboratory nature also became relatively common on the service bench. Actually, the service technician had become of age.

In industry, which was quickly converting to the television program, the service situation was given considerable attention. This attention, however, was concentrated on the sales end, and many engineers drifted back to a state of indifference. Although design layout is probably better in most cases than prior to the war, very little serious consideration was given to service problems. One major manufacturer, who had been well educated during the war, was very careful to provide test points, but this is the exception rather than the rule.

To point out the lack of service engineering in television receivers, a few of the more

* Decimal classification: R730×R583. Original manuscript received by the Institute, July 28, 1949; revised manuscript received, August 23, 1949.

† Bergen-Passaic Electronics, Inc., Bogota, N. J.

obvious deficiencies can be cited. These cases completely ignore the circuit design but deal strictly with physical layout or common-sense reasoning. One of the most glaring difficulties is that a model is designed not giving much consideration to the final relationship between the cabinet and the chassis. A common case of this type is the problem encountered in removing the protective can around the high-voltage circuit on several current models of receivers. In order to remove this can, the entire chassis must be removed from the set. A simple redesign would allow the serviceman access, saving some ten or fifteen minutes of labor. One popular model (Fig. 1(a) and (b)) has

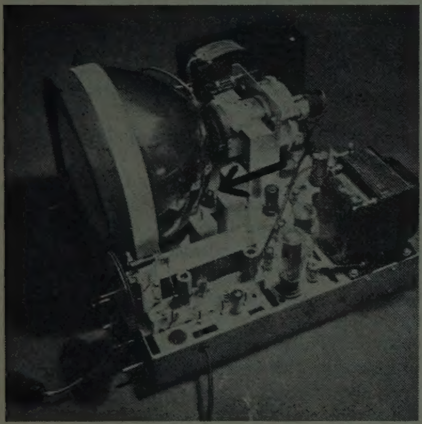


Fig. 1(a)—Receiving-type tube shown by arrow is hard to remove because of lack of room for the hand, and cannot be replaced when set is in cabinet.

Fig. 1(b)—Production changes resulted in tube shown in Fig. 1(a) to be relocated on bottom side of chassis.

receiving-type tubes located both underneath the cathode-ray tube and underneath the chassis, necessitating removal of the chassis from the cabinet to try a simple tube replacement. This same receiver has two different types of rf tuners electrically interchangeable but which are not mechanically

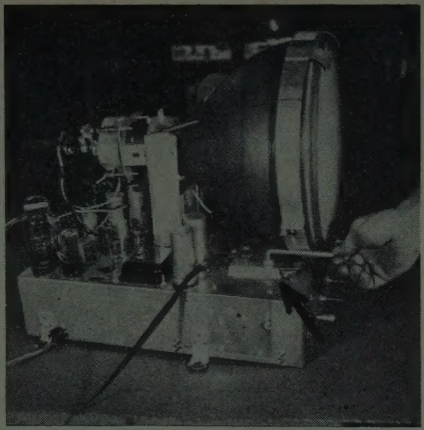


Fig. 2(a)—Tubes in "front end" of this set covered by plate, necessitating removal of chassis from cabinet and use of offset or end wrench.

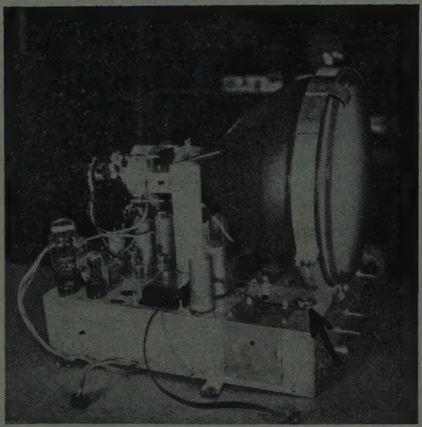


Fig. 2(b)—When plate is removed from set shown in Fig. 2(a), physical location does not allow finger space required for removal of tight-fitting shield.

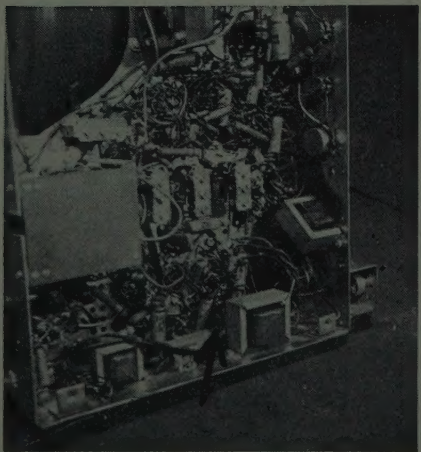


Fig. 3—Tube socket and several components lack accessibility due to placement of choke.

interchangeable on all sets simply because of the location of the screw holes. A model of another manufacturer requires removal of the chassis from the receiver to tune the oscillators in the rf tuner (Fig. 2(a) and (b)). In this model it is also extremely difficult to replace the oscillator tube because insufficient room is allowed for the serviceman's fingers. Still another manufacturer has a model on the market in which it is not possible to check the operating voltages on the converter tube because the channel-selector switch is in the way (Fig. 4). On another make of slightly older vintage, the manufacturer has the audio-output transformer conveniently riveted to the chassis in such a position that the socket to one of the tubes is very neatly protected from a soldering iron and voltmeter probe (Fig. 3). Another quality receiver has utilized three different rf tuners in the same model, each of which requires other circuit differences.

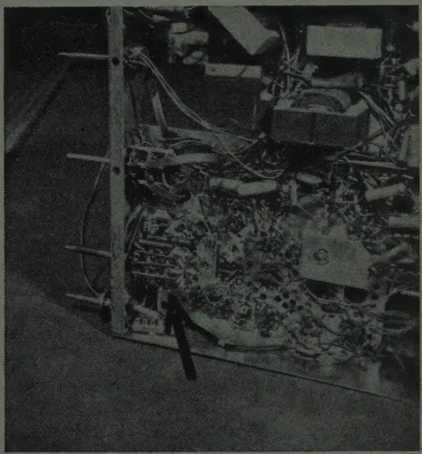


Fig. 4—Station selector assemblies are notably difficult to service. Tube-socket pin connections are inaccessible in this model. Oscillator coils are nonadjustable, necessitating use of heat and movement of windings to adjust frequency.

There are probably many methods of approach to the problem, depending upon the organization of the manufacturing firm and the personnel involved. Each organization must determine its own method of approach. In some cases, a cooperative spirit between the engineering and service departments will suffice. In other cases, it may be necessary to assign a service engineer to the engineering section and charge him with full responsibility in order to avoid the feeling that an "outsider" is attempting to advise the engineering department of their responsibilities. Regardless of the approach, there is no question but that there is a big job to be done and that it can be done in a satisfactory manner. The experiences of the Armed Forces, as previously pointed out, have fully well demonstrated the possibilities of such work. The fact that a project of this nature has been so well handled by the government should create a ringing challenge to industry. Let us hope that industry will respond as it has so many times in the past.

Adjustment of High-Precision Frequency and Time Standards*

JOHN M. SHAULL†

Summary—**High-precision frequency and time standards are becoming more and more widely used in many technical fields. The basic equipment used by the Central Radio Propagation Laboratory of the National Bureau of Standards in providing and broadcasting standard frequency and time signals is discussed. Mention is made of the manner in which it is adjusted and the corrections which may be applied to make use of the ultimate accuracy that may be expected in using these signals for measuring and calibrating similar equipment. Several methods are given for checking the frequency of precision oscillators and precision clock performance. Suggestions are given as to methods of recording and evaluating performance data for such standards. Expected improvements in constancy and accuracy and possible future changes in the types of standards used in physical time measurements are considered.**

INTRODUCTION

HIGH-PRECISION frequency and time standards are becoming more and more widely used in technical and scientific fields. Among their uses are: control sources for standard-frequency and time broadcasting; time standards for astronomical observatories; synchronization of master and slave stations for pulse navigation systems; investigation of long-distance radio transmission phenomena by pulse techniques; as the heart of frequency synthesizers in large communications systems; and in physical research laboratories for precise calibrations and measurements, such as time-rate phenomena and microwave spectroscopy.

While this paper deals primarily with the adjustment and performance of high-precision frequency and time standards, many of the principles involved and techniques discussed should prove helpful to those using the WWV frequencies or time signals in adjusting or calibrating other high or medium precision equipment.

The accuracies presently required by some users of frequency and time services are approximately as shown in Table I.

TABLE I

Service	Accuracy (frequency)	Sec/day (time)
Physical research Astronomers	1×10^{-8}	±0.001
Monitoring stations Surveyors	1×10^{-7}	±0.01
Radio broadcasting Astro-navigators	1×10^{-6}	±0.1
Commercial communication	1×10^{-5}	
Musical instruments	1×10^{-4}	
Commercial power distribution	1×10^{-3}	(±5)

* Decimal classification: R214. Original manuscript received by the Institute, September 27, 1949.

† National Bureau of Standards, Washington, D. C.

Equipment and Methods in General Use

A high-precision frequency standard may be defined as one whose changes in frequency are less than 1 part in 10^8 per day. This degree of precision now requires continuous operation of the standard oscillator, with the control unit and other critical elements operating in a temperature-stabilized compartment. Very careful shielding, filtering of battery supply leads, and use of buffer amplifier stages are also necessary, especially if more than one oscillator unit is operated at a single location. Such standards now generally employ a GT-cut quartz-crystal[1] operated in a bridge-stabilized oscillator circuit arrangement.[2]

A standard oscillator in wide use operates at 100 kc and has a multiposition switch for coarse frequency adjustments of approximately 4 parts in 10^6 per step. A precision gear-driven capacitor with a drum dial arrangement of 5,000 dial divisions provides for a control of frequency to approximately 1 part in 10^9 per division. A dial with such an expanded scale is of great advantage in making precise frequency adjustments and interpolations.

A minimum of three standard oscillators is recommended for those installations requiring a reasonable maximum of reliability and continuity of service. By intercomparing three standards locally, either once daily or continuously, short-time stability may be determined to a much higher order than possible through radio transmission comparisons. The most reliable standard may thus be determined and used to supply the desired need, with the other units available for standby duty.

For checking frequencies by the time-comparison method, and for supplying the desired audio frequencies and time intervals, each standard oscillator should be operated continuously, and two or more such oscillators should be provided with frequency-dividing and synchronous-clock equipments. The frequency dividers may be pulse counters, fractional frequency generators, locked oscillators, or multivibrators. The choice of type should be influenced by the number of output frequencies and wave form desired, simplicity of adjustment, and reliability of operation required. If seconds intervals are desired for laboratory use, a system similar to that used at WWV, wherein the pulse trains are generated electronically and selected mechanically by a rotating cam, offers certain advantages. An alternate method is to divide the standard frequency down to 1 cps by counter methods, employing an all-electronic system.

A continuously adjustable phase shifter of the polyphase electrostatic or electromagnetic type is desirable in the divider chain, preferably at the 1,000-cycle stage, to permit the clock or seconds signals to be synchronized. This device also provides a convenient means of measuring small daily differences in time kept by different standard-oscillator clocks, including time signals received by radio. Measurements of time differences may be made by adjusting the phase shifter dial until the pulses coincide on an oscilloscope screen when alternately or simultaneously connected.

Other methods for measuring and recording time differences may employ a spark chronograph, or a polyphase modulator and integrating phasemeter.[3]

A harmonic generator and mixer unit may be conveniently used in rapid frequency intercomparison. This device may be provided with two inputs and means to control the inputs which feed into the crystal diode harmonic generators. The common output is connected to a radio receiver for counting the difference beats between two local standards. Two frequency multiplier units having outputs of 500 and 2,500 kc or other convenient frequencies, are recommended to facilitate frequency measurements at higher harmonics.

Standard Frequency and Time Broadcasts from WWV

The radio and audio frequencies as transmitted from WWV, near Washington, D. C.[4] are accurate (with reference to mean solar time) within 1 part in 50 million. The time signals broadcast by WWV are maintained in agreement with U. S. Naval Observatory time within several hundredths of a second. This is done by setting the WWV control oscillator frequency slightly higher or lower than exactly 100 kc by an amount ordinarily not greater than 1 part in 10^8, to advance or retard gradually the broadcast time. For this reason the time broadcast by WWV is uniform, changing by less than 0.001-second average or 0.002-second maximum per day. Fig. 1 shows the frequency and time deviations of the WWV transmissions for the years 1947 and 1948.

The present oscillators at WWV have a gradual and fairly constant drift to a higher frequency of from 0.6 to 1.2 parts in 10^9 per day. The control standard at WWV was adjusted on an average of every twelve days during the past year. Adjustments are usually to a lower frequency but are occasionally made to a higher frequency because of slight discrepancies in the assessment of average drift rate in terms of the Observatory's time determinations. The adjustments

[1] W. P. Mason, "A new quartz crystal plate, designated the *GT*, which produces a very constant frequency over a wide temperature range," PROC. I.R.E., vol. 28, pp. 220–223; May, 1940.

[2] L. A. Meacham, "The bridge stabilized oscillator," PROC. I.R.E., vol. 26, pp. 1278–1294; October, 1938.

[3] W. A. Marrison, "Evolution of quartz crystal clock," *Bell Sys. Tech. Jour.*, vol. 27, pp. 510–588; July, 1948.

[4] "Technical radio broadcast services, radio station WWV," Letter circular LC886, obtainable from National Bureau of Standards, Washington 25, D. C.

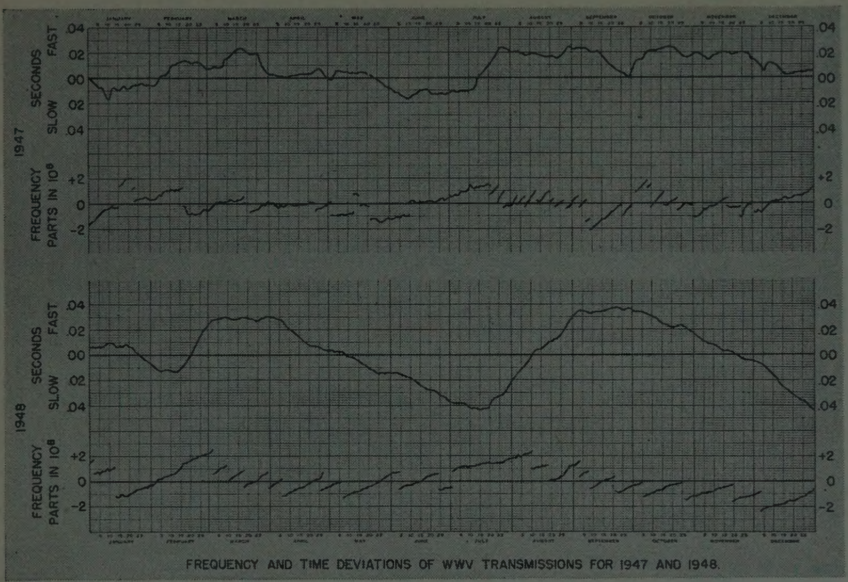


Fig. 1—Frequency and time deviations of WWV transmissions for 1947 and 1948.

are usually made on Fridays between 9 and 10 A.M. EST. but may be made on any day of the week.

A second standard-frequency station WWVH, located in Maui, T. H., has been operating experimentally since November, 1948. The purposes of this station are to learn ways of improving coverage in areas not well served by WWV and to determine the feasibility of operating several standard-frequency stations in different parts of the world on the same frequencies. At present, WWVH provides transmissions on 5, 10, and 15 Mc with carrier powers of 400 watts for each frequency. The carrier frequencies and modulation components are derived from primary frequency standards similar to those at WWV. These oscillators are maintained in agreement with the WWV transmitted frequencies within 2 parts in 10^8 by means of time comparisons over 6-day periods.

The National Primary Standard of Frequency

The present primary standard of frequency consists of three precision frequency standards located at the WWV transmitting site near Greenbelt, Md., and five located at the National Bureau of Standards in the District of Columbia. These eight standards are automatically compared with each other and with Naval Observatory time determinations. The estimated or daily assigned frequencies, accurate to about 1 part in 10^8, are determined by a weighted extrapolation process in terms of the most reliable standards. Corrected frequencies are evaluated, over 100-day intervals, in a manner explained under "Interval-derived frequencies" and "Long-interval performance determinations" in terms of the Naval Observatory time signals for three well-stabilized standard oscillators. The mean of these corrected values applied through the daily beat-frequency comparisons is adopted as the final daily corrected or "absolute" frequency for each standard. This value is obtained some 60 days later and is used as a guide in extrapolating the daily assigned frequencies so as to keep their corrections as small as possible. Agreement of the 100-day corrected frequencies computed separately in terms of each of the three interval-derived curves is generally within 2 parts in 10^9. The principal limitation on the obtainable accuracy is thus evidently caused by the slight wanderings in the determinations of the earth's mean rate of rotation.

Daily and short-time variations in the individual primary frequency standards are of the order of 1 or 2 parts in 10^9 and 1 or 2 parts in 10^{10}, respectively. Fig. 2 is a 24-hour chart recording of the beat frequencies a

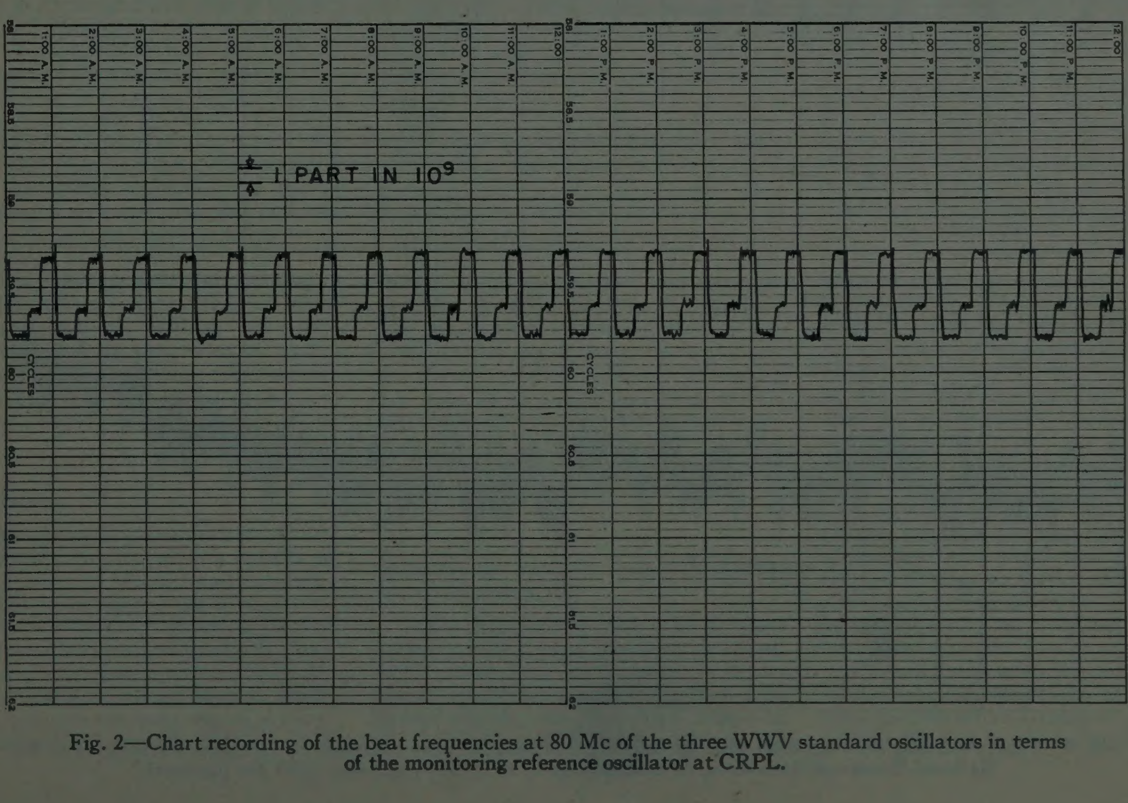


Fig. 2—Chart recording of the beat frequencies at 80 Mc of the three WWV standard oscillators in terms of the monitoring reference oscillator at CRPL.

80 Mc of the three WWV oscillators in terms of the monitoring reference standard, showing the order of these variations. Fig. 3 is a chart recording showing short-time variations in the beat frequency between two standards, on different days, at 80 and 1,280 Mc. The small variations of about 2 parts in 10^{10} are attributable to slight mutual coupling between the units and to heater thermostat operations. Fig. 4 shows chart plottings of the three WWV standards, primary 25, primary W1, primary 65, and the two monitoring reference standards, primary K15 and primary 35, for several months.

Measurement of Precision Standard Oscillators by Frequency Comparisons with WWV

For direct frequency measurements, a harmonic of the local frequency standard is compared with the received WWV signal to determine the difference frequency. Where several standard oscillators are available and regularly intercompared, it is necessary to check only one in terms of WWV. Harmonic power from the local standard to the radio receiver should be adjusted to be about equal to the received signal level so as to obtain a maximum modulation or beat. During severe fading, it may be necessary to count beats that are suppressed but would continue the natural rhythm of those observed. When reception is good, the best results should be obtained by counting beats over a 1- to 2-minute period. When fading is severe, the averaging of a larger number of successful counts of periods of from 10 to 20 seconds may prove most useful.

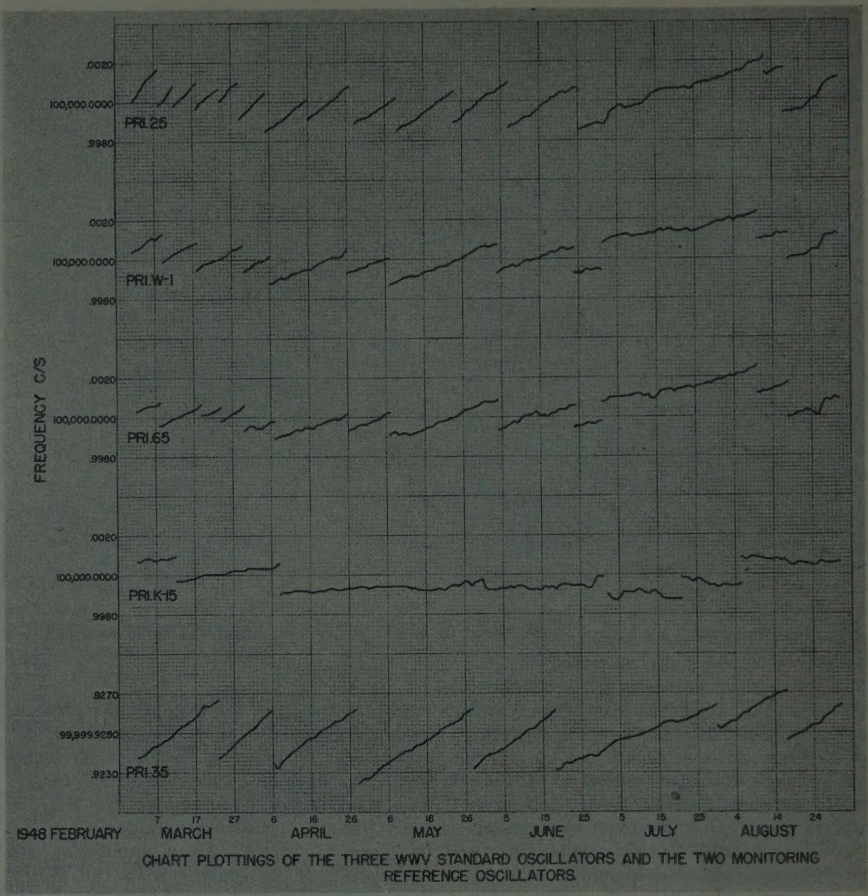


Fig. 4—Chart plottings of the three WWV standard oscillators and the two monitoring reference oscillators.

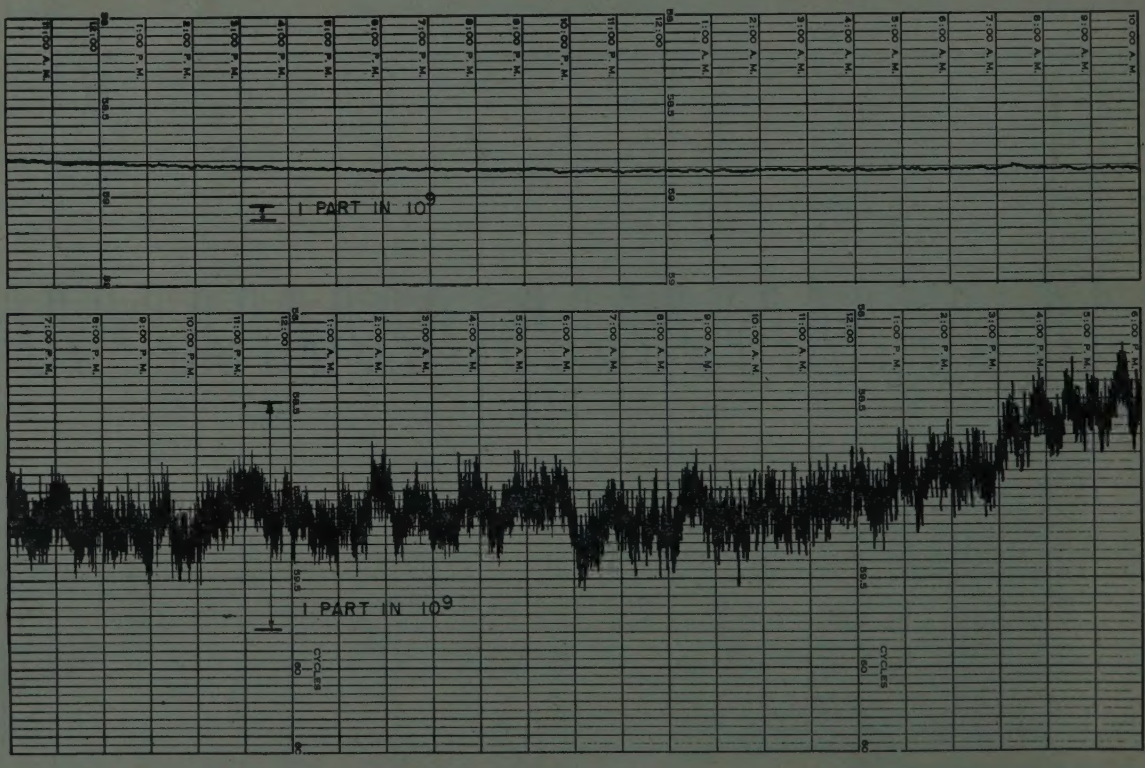


Fig. 3—Chart recording showing short-time variations in the beat frequency between two frequency standards at the National Bureau of Standards, on different days, at 80 Mc (top) and 1,280 Mc (bottom).

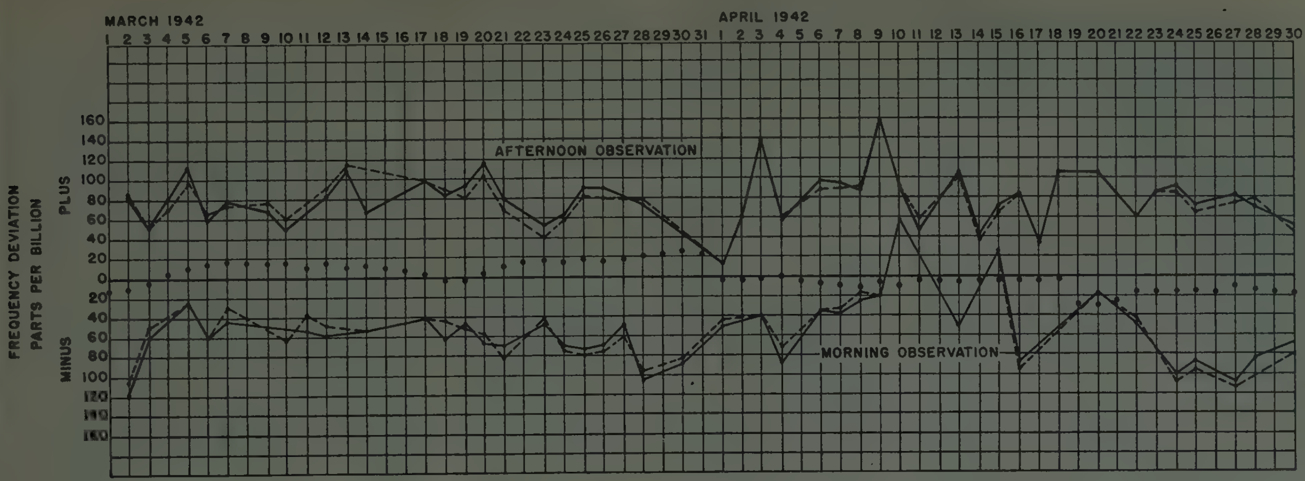


Fig. 5—Illustration of Doppler variation of the WWV received frequencies in terms of two frequency standards at Western Electric's Hawthorne, Chicago, Laboratory. Dotted points are WWV transmitted frequencies.

Experience has shown that it is often quite difficult to adjust a local compared oscillator directly to zero beat frequency with WWV at remote locations by either aural or S-meter means. This is generally caused by numerous rapid variations in signal level, or by propagational flutter in the received frequency. Tests at the National Bureau of Standards where a ground-wave signal from WWV prevails, showed no difficulty in reproducing zero-beat settings of an oscillator dial to ± 1 part in 10^9.

Multiple-Setting Interpolation Method

Multiple-setting interpolation is best carried out by setting the fine-frequency dial of the compared oscillator slightly off frequency so as to obtain beats with WWV of from 0.1 to 2 beats per second. Several settings should be made on each side of zero beat, choosing values which are readily distinguished from fading, and a number of beat periods measured by means of a stop watch. Values thus obtained are then converted to beat frequencies in cycles per second and are plotted on rectangular co-ordinate paper as ordinates with dial settings as abscissa. By plotting the beat frequencies obtained below zero beat negative and the ones above zero beat positive, and drawing a straight line through the average of these points, an accurate zero beat dial setting will be indicated where this line crosses the x axis. For this method an adjustment dial having good linearity, very little backlash and one dial division equal to about 1 part in 10^9 is required.

Offset Method

The multiple-setting method is somewhat cumbersome in that it requires a number of separate settings of the dial and computations and involves plotting the data, preferably on rather large graph paper. As the dial setting versus frequency is generally quite linear over a small frequency range, the desired zero-beat setting may be more easily obtained to about the same degree of accuracy by using an offset method. This is done by taking the average of several counts of beats at a single offset point at approximately 1 part in 10^7 from the zero-beat point. The average of these counts in seconds per beat is converted to beats per second which represents the amount the oscillator is then off zero beat at the standard frequency to which the receiver is tuned. This is then converted to parts-in-10^9 deviation from zero beat by dividing the beat in cycles per second by the standard frequency in cycles per second and multiplying by 10^9. To return the oscillator to zero beat it is only necessary to advance or retard the dial the indicated number of parts in 10^9 with allowance for the incremental dial correction factor for the region in which it is operating. The dial correction factor (number of parts in 10^9 per dial division) is most readily obtained by comparing the standard with a similar local standard at several dial settings, and thus avoiding radio transmission difficulties and possible errors.

This method has the advantage of requiring only one offset adjustment. With either method the amount of offset should be made small as any adjustments may temporarily disturb the normal drift rate to some extent, even though the dial is returned to the original setting. If several units are available in the frequency standard arrangement, the oscillator which is off frequency by a small amount and infrequently adjusted may be used to make the comparison with the transmitted frequency. The others may then be evaluated in terms of this standard so that no changes in the others are necessary. The offset standard should be sufficiently off-frequency so as to be certain that it is higher or lower than the received frequency.

Variations in Received Frequency Arising from the Doppler Effect

The accuracy of any method using direct comparison of a local frequency standard with a received standard frequency will be affected by fluctuations in transit time of the received signals. This arises from changes in the radio propagation medium during the measurement period, which make the received frequency slightly lower or higher than that transmitted. This well-established Doppler effect could be readily computed for radio propagation if sufficient information on changes in the medium were available. Its bearing on frequency measurements may be outlined as follows. The ionospheric reflecting layers vary in height considerably with frequency, time of day, season, geographical location, and phase of sunspot cycle. Average conditions are predictable, but conditions on a given day may depart greatly from the mean. Changes in effective layer height will cause the frequency of a received carrier to differ from that transmitted by a fractional amount equal to the rate of change in equivalent path length on kilometers per second divided by the propagation velocity in kilometers per second. For example, if one assumes $F2$-layer propagation and that the virtual height where reflections occur changes from 300 to 400 km in 2 hours, then for a "3-hop" transmission from WWV, Washington, to WWVH, Maui, (7,700-km surface distance) an effective change in frequency of the received signal of -3.2 parts in 10^8 for each hop can be computed. On a similar basis the computation for "4-hop" transmission (at a higher angle of departure) gives -3.7 parts in 10^8 per hop so affected. Of course, it is unlikely that changes of this magnitude will occur simultaneously at all reflection points of the path considered; thus the error caused by the Doppler effect will usually be less than the above figures multiplied by the number of hops.

The WWV frequencies as received at Maui, T. H., in October, 1948, were generally slightly higher in the morning and slightly lower in the afternoon. Measurements based on reception at a time when noon occurred about halfway between transmitter and receiver showed consistent agreement to about 1 part in 10^8, with occasional discrepancies as great as 2 or 3 parts in 10^8.

Observations made in England on WWV's 15-Mc received frequencies in December, 1945, showed errors ranging from 2 to -7 parts in 10^8, with an average variation of -3

parts in 10^8 when compared with very stable primary frequency standards.[5] Similar measurements reported by Booth and Gregory showed slightly greater variations.[6] Fig. 5 shows the variation in WWV's 5-Mc received frequency at Western Electric's Hawthorne, Chicago, Laboratory in terms of two precision frequency standards. It is evident that high-accuracy measurements of frequency in terms of WWV should be made when ionospheric layer heights are likely to be most stable, i.e., with noon or midnight prevailing at about halfway between transmitter and receiver locations. Long-distance north-south transmission may be expected to show greater variations in the morning and afternoon where all of the reflection points are subject to changing conditions at the same time. Long-distance comparisons should, in most cases, prove most reliable when lower angles and fewer modes of propagation prevail. However, one should particularly avoid use of a frequency and time where a dominant mode of transmission is very near the maximum usable frequency at the propagation angle for any of the reflection points, as the effective layer height changes very rapidly under these conditions. It should be mentioned that errors other than those attributed to the Doppler effect were occasionally found possible.

Publications of the National Bureau of Standards are available which are useful in determining the optimum reception frequencies at a given time and location.[7,8]

Measurement of Precision Standard Oscillators by Daily Time Comparisons with WWV

The average frequency of a standard oscillator may be determined by successively comparing the number of cycles generated (time kept by its synchronously-operated clock) with any sufficiently reliable time signals. If the standard's drift rate or gradual change in frequency is assumed to be constant, its average frequency during the period will be the same as the instantaneous frequency for the center of the period considered. Departures from a constant or uniform frequency drift can be detected by this method only by successive determinations of frequency, and only then if the reference time source is known to be extremely accurate. When using time signals transmitted by radio over long distances (sky-wave propagation), slight errors will result because of variations in the time of transmission of as much as several milliseconds under adverse conditions. For this reason frequency determinations based on time comparisons of less than about two days may be in greater error than direct-frequency comparisons with transmitted standard frequencies.

Interval-Derived Frequency Determinations

The computation of average frequency by time comparisons may be explained by the following equations:

$$f_{av} = f_0 \frac{t_2 - t_1}{T_2 - T_1} \qquad (1)$$

where f_0 is the nominal frequency of the oscillator, i.e., that frequency for which the frequency division ratios were chosen; t_1, t_2 and T_1, T_2 represent the time indicated by the clock, and the correct time represented by the time signals respectively at the beginning and end of the period being averaged.

Equation (1) is usually simplified in actual use by referring all measurements to days or multiples thereof and by using measured errors in time indicated by a crystal clock. For example, the average frequency of a 100-kc standard oscillator during interval T is

$$f_{av} = 10^5 \left(1 + \frac{\Delta t_2 - \Delta t_1}{T}\right), \qquad (2)$$

where Δt_1 and Δt_2 are positive or negative errors in time at beginning and end of the measurement period. The value obtained represents the instantaneous frequency for the center of the period, if uniform drift is assumed, or the value for the entire period if no change in frequency during the entire period is assumed. Instantaneous departures from the average values are generally determined by daily comparisons with other oscillators, involving an averaging based on their predicted daily drifts. This is justified by the probability that the average of the drifts of a number of selected oscillators will be more nearly linear than the drift of any one unit.

Daily measurements of time differences or errors between standard clocks and time signals may be accomplished by adjustment of a calibrated phase shifter dial for pulse coincidence on an oscilloscope screen or by use of calibrated circular or linear sweeps on the oscilloscope. A recording chronograph may also be used to get a continuous record of the time differences of several standard clocks and time signals.

Clock Acceleration

In order to determine the amount of time a crystal clock will gain or lose over a long period it is necessary to consider the effect of the change in frequency (and thus the change in rate of the clock) over the period. For very long intervals it is even necessary to consider the change in the drift rate of the oscillator (change in the change of rate of the clock).

The time a crystal clock will indicate at some future time T_1 at t days distant is:

$$T_1 = T_0 + t + at + bt^2 + ct^3 \qquad (3)$$

where at represents the "rate" term, bt_2 the acceleration term, and ct^3 the change in acceleration term.

The terms T_0 and t are always positive. Coefficient 'a' may be positive or negative, and is the principal factor in determining the overall rate of the clock over short periods. Coefficient 'b' is generally positive, and must be considered for periods of more than a few days. Coefficient 'c' is very small and is generally negative. It may be neglected in all but very long-time computations.

The above equation may be more readily understood if times T_0, T_1, T_2, etc., are considered as marking off an absolute, continuous time scale, with the expression $t+at+bt^2+ct^3$ marking off the distance covered by the clock on this time scale in t days. This may be recognized as the well-known equation for linear motion. Usually $t+at$ is combined and considered as the velocity term, but for horological purposes it is more convenient to keep these two terms separate. Horologists express the "rate" of a clock as the seconds gained or lost per day.

Fig. 6 shows the computed daily change in time of a crystal clock for a uniform daily change in frequency of 1 part in 10^9 per day. The clocks are usually set slightly fast in such manner as to lose time, and become slightly slow as the oscillator passes through correct frequency. As the oscillator becomes high in frequency, this time is regained to approximately the original setting when the oscillator frequency is about 1 part in 10^7 high. The oscillator is then adjusted low again and the cycle repeated, this adjust-

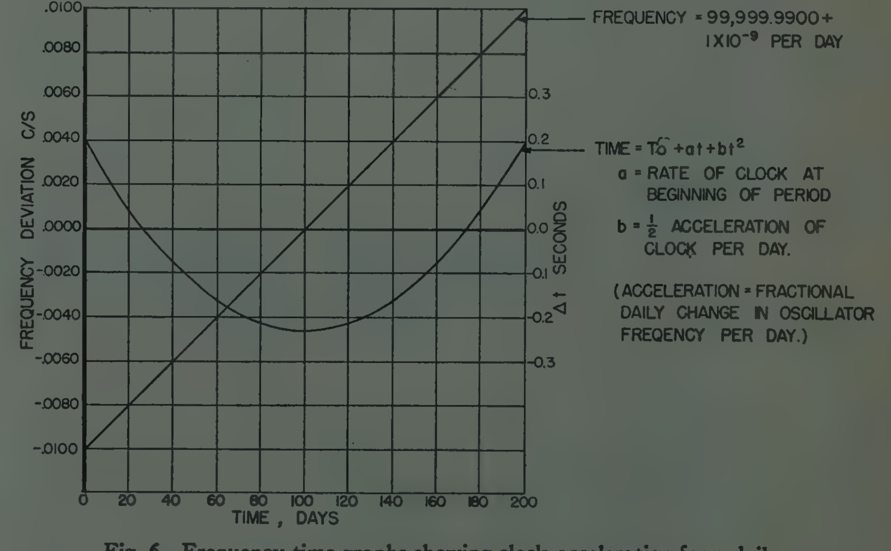


Fig. 6—Frequency-time graphs showing clock acceleration for a daily change of frequency of +1 part in 10^9.

[5] H. V. Griffiths, "Doppler effect in propagation," *Wireless Eng.*, vol. 24, pp. 162–167; June, 1947.

[6] C. F. Booth and G. Gregory, "The effect of Doppler's principle on the comparison of standard frequencies over a transatlantic radio path," *Post Office Elec. Eng. Jour.*, vol. 40, pp. 153–158; January, 1948.

[7] "Ionospheric radio propagation," Circular 462, issued June 25, 1948, available from Superintendent of Documents, Government Printing Office, Washington 25, D. C. (price $1.00; foreign $1.25).

[8] "Basic radio propagation predictions," CRPL-D series (monthly, three months in advance) available on subscription (price $1.00 yearly, foreign $1.25) from Superintendent of Documents, Government Printing Office, Washington 25, D. C.

ment being required about once or twice per year to maintain frequency limits of ±1 part in 10^7.

Propagation Delay of Time Signals

Changes in the height of the ionosphere layers by which time signals or seconds pulses are received, or propagation by different layers or numbers of hops on successive measurement periods will introduce variable delay in the time of reception. The amount of time delay introduced by transmission time may be computed if the mode of transmission is known or assumed. This delay is nearly 1 millisecond per 186 miles of actual transmission path length. The difference in reception time for signals propagated by various modes and going only one way around the earth is seldom greater than a few milliseconds. By observing the earliest consistently received pulses over a short period, this may be generally reduced to less than 1 millisecond.

As an illustration of the relative delays to be expected from various modes of propagation, the time lags for a curved earth-ionosphere path for several propagation conditions have been computed. These values for transmission from WWV, Washington, to WWVH, Maui, a great-circle ground-path length of about 7,700 km (4,800 mi) are as shown in Table I. Geometrical one-hop transmission over this path is impossible, and two-hop transmission would be expected to occur less frequently than transmission by the higher modes for this path length. If $F1$-layer transmission is assumed the delay should be generally a little less than for the similar $F2$-layer mode. These values assume propagation by a number of geometrical hops and normal reflections at the indicated virtual heights. Scattering, interlayer reflections, ionospheric turbulence, and abnormal conditions are of frequent occurrence.

TABLE I

No. of hops	F layer (210 mi.)	F_2 layer (267 mi.)
	Milliseconds delay	Milliseconds delay
2	26.6	27.2
3	27.0	27.7
4	27.9	29.1
0 ground path equivalent lag (for comparison purposes) is 25.7		

The average variations in delay of the WWV seconds pulses as received at Maui, Hawaii, over a number of one-hour observation periods ranged from 0.1 to 0.2 millisecond for observations on a single frequency, and 0.6 to 1.0 millisecond for observations on the several frequencies receivable at a given time. These variations represent the deviations in a number of loggings of the earliest pulses consistently received over short intervals and not the scatter of multiple pulses received at the same time or consecutively. Variation of transmission delay time of about 2 milliseconds has been reported on high-frequency time signals over the North Atlantic path.[9]

[9] C. F. Booth and F. J. M. Laver, "A standard of frequency and its applications," *Jour. IEE*, vol. 93, part III, pp. 223–241; July, 1946.

$$\text{beat } C/S = \frac{\text{number of beats counted}}{\text{number of seconds} \times \text{receiver frequency in Mc} \times 10^{8}}.$$

When using the WWV or WWVH seconds pulses at locations where both stations are received, one or the other signal should be selected when making time measurements for frequency determination. This may be done by using a directive receiving antenna or by selecting the proper pulses when making observations. Similarly, at certain locations it is necessary to differentiate between pulses received from both ways around the earth. For example, at Maui pulses are consistently received in the morning from WWV on 15 and 20 Mc by paths going both ways around the earth, with delays of approximately 0.027 and 0.113 second. Quite often, for brief periods, the pulses received over the longer path are stronger than those received over the shorter path because of differences in absorption. When checking a well-stabilized standard clock, the expected time of arrival will be known to a few milliseconds, which helps to discriminate against undesired signals. As with the frequency comparisons, it is desirable to make these measurements when reception conditions are optimum and the ionosphere is stable.

Interval-derived frequency determinations covering periods of 1 or 2 days may not prove much more accurate than those made by direct frequency comparisons under optimum conditions. However, if highly stable oscillators are considered, and the measurement period is increased to about 6 to 10 days, comparison accuracies of a few parts in 10^9 may be obtained.

Intercomparison of Local Standards

When more than one frequency standard is available, it is generally desirable to check each local standard at regular intervals or continuously with the local reference oscillator (one checked in terms of WWV). The beat frequencies between the reference and each of the other oscillators are then determined for plotting purposes at the time chosen for evaluating the reference oscillator. This may be done by one of the following methods.

Local Frequency Intercomparison by Use of Harmonic Mixer and Receiver

The harmonic mixer unit has already been mentioned in connection with WWV frequency comparisons. Measurements are made by adjusting the 100-kc input of each source separately by means of the level controls to about S-4 on the radio receiver. Both oscillators are then connected, and the beat counted on the S-meter at any convenient frequency multiple of 100 kc. To get the desired accuracy the beats should be counted using a stop watch for a minimum of 2 beats, or a convenient even number of beats which can be counted in about 2 minutes. The difference frequency in cycles per second at 100 kc is then

Local Frequency Intercomparison by Use of Oscilloscope

The beat frequency between two 100-kc frequency standards may be readily obtained to a high degree of precision by observing high-order Lissajous patterns on an oscilloscope. This may be conveniently done by connecting the output from a frequency multiplier controlled by one standard to the vertical deflection plates or their wideband amplifier input, and 100 kc from the other standard through the horizontal amplifier to the horizontal deflection plates. The horizontal pattern should be expanded to get sufficient resolution of the multiple pattern. Phase changes, caused by any difference in frequency, will cause the pattern to close and open in continuous sequence. A point near the center of the screen may be chosen where the lines close and the time for an even number of closures measured with a stop watch.

The difference frequency in parts in 10^9 is:

$$\text{Frequency difference} = \frac{500 \times \text{number of closures counted}}{\text{time of count in seconds} \times \text{highest frequency used in Mc}}.$$

A minimum of two closures or a minimum counting time of at least two minutes should be used for the desired accuracy. These observations will not indicate which of the two frequencies is higher. However, by using the frequency from one oscillator to lock in the linear-sweep oscillator in the oscilloscope, observation of the direction of the pattern drift will indicate which of the two frequencies is higher.

Other Precision Frequency Intercomparison Methods

For a complete installation of standard oscillators, some method should be provided for continuously and automatically recording the difference or beat frequencies between the reference oscillator and the others. Several methods will be described, although others may be employed. The direct-reading, or easily computable sensitivity, should be 1 part in 10^9 or better. The maximum deviation capability of the recorder need not be greater than a few parts in 10^8. The instrument should be capable of indicating short-interval stability conforming to a sampling or reading completed in a minute or less time.

One very satisfactory but somewhat elaborate method, which is direct-reading and unambiguous with respect to beat sign, uses a commercial power-frequency 60-cycle recorder to record the beat frequencies at 100 Mc. Two 100-kc to 100-Mc frequency multipliers are used. A continuously adjustable phase shifter of the rotating or electronic type is used to subtract exactly 60 cps (obtained from the standard) from the output of the reference multiplier at 100 Mc. The resultant frequency and the output from the other multiplier are supplied to a converter and the difference frequency of approximately 60 cps obtained. The difference fre-

quency is amplified and supplied to the 58 to 62 cps frequency recorder so that the 60-cycle point represents zero difference frequency. This gives a continuous record of the difference frequency to a sensitivity of a few parts in 10^{10} with a range of ± 2 parts in 10^8. The other oscillators can be switched on successively in rotation to record the deviation of each one in terms of the reference oscillator.

A variation of this method which is currently used at National Bureau of Standards is to set the reference oscillator approximately 60 parts low in 100 million, which gives a recordable difference frequency without the use of the phase-shifting mechanism. This method has the disadvantage of putting the reference off frequency by an amount making it unusable for many purposes.

The difference frequency between two standard oscillators may be determined by accurately measuring the time required to complete one beat at the fundamental or a multiplied harmonic of each standard. A method for doing this has been described by H. B. Law, and is used by the British Post Office and the National Physical Laboratory for the comparison of precision frequency standards.[10] A balanced phase discriminator is supplied with two 100-kc frequencies and arranged to trigger a counter chronometer at the beginning and ending of one beat. The chronometer counts a convenient standard frequency (100 or 10 kc) and thus indicates the elapsed time of one beat. The instrument accuracy under ideal conditions is estimated to be 1 part in 10^{11}.

Two electronic counter chronometers may be used in a number of ways to obtain accurate comparison data between standard oscillators. A general method uses one instrument to count the difference or beat frequency between high harmonics of the two compared frequencies, while a second instrument counts cycles of appropriate standard frequency. One unit may be adjusted to serve as a predetermined counter to stop the other unit after a definite count and thus increase the accuracy somewhat. This method of frequency measurement is very flexible and can be used to measure frequencies which fall outside the limits of other very sensitive measurement devices.

A very simple but accurate method of recording low beat frequencies such as obtained from the harmonic mixer and receiver comparison method is to use a recording milliammeter or recording oscillograph. The beats are then evaluated in terms of uniform chart speed or time markers recorded simultaneously. This method offers comparative simplicity for temporary or experimental use, but is quite laborious where a number of readings are required over a long period of time.

Recording and Evaluation of Data

The amount of statistical data taken on each standard should be kept to the minimum necessary to determine the probable performance of each individual unit over a continuous period. The chief reason for taking such data on a number of standards is to determine the most suitable one for use, and to determine its short-time stability in terms of the other available standards. When such a standard is used to control standard-frequency transmissions or to make high-precision calibrations, it should be monitored continuously against one or more similar standards.

As a result of experience gained in the operation of frequency-standard oscillators at the National Bureau of Standards and in particular at the WWVH, Maui, T. H., transmitting station, it may be helpful to outline maintenance, procedures and techniques that have been developed and found useful in this work.

Logging and Plotting of Data

At the NBS it has been found necessary to keep a daily record which includes all frequency and time readings and measurements, as well as the final results of all computations of frequency and time. Changes in control standard, methods, or equipment are always noted. These data are conveniently kept in a record book having a number of ruled columns, or on specially prepared mineographed sheets carrying the desired notations for a particular installation.

The daily computed frequencies for each oscillator are plotted on continuous cross-section paper in a manner as shown in Fig. 4. These data are of considerable aid in studying the relative performance of the individual standards, and in graphically extrapolating predicted values during periods when reception conditions do not permit making the daily frequency or time checks.

It has been found desirable to keep periodic (weekly) instrument readings which include the readings of all battery chargers and power supplies, voltmeter readings on each switch position for the frequency standards and frequency dividers, temperature readings of each standard's oven where applicable, and room temperature at time of readings. These readings are useful in anticipating tube or equipment failure in that adjustment or repair can often be made before complete breakdown occurs.

Adjustments in Frequency and Time

Periodic adjustments in frequency and time of each standard oscillator and time equipment are required to keep within the desired accuracy tolerance and to simplify the measuring and plotting requirements. The frequency and range of these adjustments are influenced considerably by the type of service and standard. For greatest constancy of drift rate and greatest ease in computation, frequency adjustments should be held to a minimum. For many uses the calculation of and allowance for the slight error in frequency or time presents no great difficulty. In other cases, where numerous measurements are being made and services given, it is helpful to establish high initial accuracy so as to eliminate corrections.

Where the working standard (one used or distributed) is held to very close accuracy tolerance, the continuity of service required will determine if the second or standby standard need be held to a similar close tolerance. If continuity of service is only of moderate importance, two equipments may be infrequently adjusted and only the working unit held within very precise limits. The oscillator and time equipment least adjusted should be generally used as the reference or one by which frequency and time determinations are made with respect to WWV. For most purposes an adjustment of the working standard so as to hold its frequency within 1 or 2 parts in 10^8 of WWV's average frequency should prove satisfactory, with the standby and spare oscillators being held to a frequency tolerance of 1 part in 10^7.

Daily phase shifter (time) adjustments may be made if time synchronization is desired, and these daily changes used in computing interval-derived frequencies. Allowance should be made for transmission time lag in setting or using the local time pulses, if extremely precise time synchronization is important.

Short-Interval Performance

High-precision standard-frequency oscillators now in use have short-period stabilities (intervals of one minute or so) ranging from 1 part in 10^9 to 1 part in 10^{10}. In addition to the gradual ageing or drift, various causes will contribute to these short-time frequency fluctuations. Mechanical shock may cause an instantaneous effect through easing of stresses or displacements in the crystal unit or associated components. A slow recovery in frequency may or may not take place. A change in supply voltage produces an immediate effect followed by additional changes as thermal equilibrium is restored. Changes in load impedance will cause frequency variations unless proper decoupling is employed. Unless proper thermal lagging is provided for the crystal compartment cyclic operations of its heaters will also cause variations in the output frequency. The oscillator circuits must be protected from direct or electromagnetic mutual coupling to better than 100 db to reduce frequency "pulling" or tendency to synchronize. Other changes in frequency, often of unpredictable source, may be caused by imperfect connections, faulty components or erratic vacuum tubes. These causes of poor frequency stability can generally be eliminated only by a slow process of substitution and observation.

Where average frequencies are computed for periods of more than a few days, the daily values may be determined by calculating the departure of the reference standard from its average curve in terms of the other reliable standards available. The daily values for the other standards may then be determined by adding the daily beat frequencies for each standard to the computed value for the reference. The mean values of relative frequencies thus obtained represent a weighted extrapolation in terms of the selected group of reliable standards. Considerations involving three independent reliable standards should give a relative daily accuracy within 2 parts in 10^9. The inclusion of six or more standards, as is done in the CRPL primary standard of frequency, results in a relative daily determination of performance of each standard oscillator to within 1 part in 10^9. Very-short period or

[10] H. B. Law, "An instrument for short-period frequency comparisons of great accuracy," *Jour. IEE* vol. 94, part III, pp. 38–41; January, 1947.

"instantaneous" stability may be determined by intercomparing pairs of standard oscillators by high-precision automatic recording equipment.

Long-Interval Performance Determinations

The determination of deviations over periods of months and years is complicated and obscured by the fact that the earth's mean rate of rotation (even after application of a number of established corrections) is not uniform. Changes in the length of the apparent day as large as 4 or 5 milliseconds, equivalent to a frequency change of about 1 part in 20 million, are believed to have occurred on several occasions within the last half-century. These changes occur at rather irregular intervals and in varying amounts and are, as yet, unpredictable. They are evidenced by an accumulation of error in the earth's observed angular position compared with theoretically predictable astronomical events.

Smaller variations, with periods of several weeks to slightly more than a year, are noted when comparing the earth's observed time with high-precision crystal clocks. Until very recently these variations were attributed almost entirely to clock discrepancies. Agreement of a number of precision clocks and technical improvements in methods of observing and recording star transits indicate that the earth's rate of rotation is subject to a number of more or less random small variations which add up to as much as several milliseconds per day at times. The probable error of a single time determination in terms of a number of star sights is believed to be not greater than several milliseconds and such errors are not cumulative. Some of the time variations are caused by rather irregular wanderings of the earth's poles by as much as 30 feet with a principal component having a period of slightly more than a year. Observational errors, caused by this variation in longitude (which would be zero at the equator) amount to as much as ± 20 and 30 milliseconds at the Washington and Greenwich observatories, respectively. These variations may be corrected by applying results of observations of latitude variation at a number of locations and computing the equivalent changes in longitude. The intricacies of time determination in terms of the earth's rate and the conversion from observed sidereal to mean solar time have been discussed by H. Spencer Jones, British Astronomer Royal.[11]

Another factor, although of no immediate concern in determining frequency, is the gradual slowing down of the earth's rate of rotation, chiefly because of tidal friction. This deceleration over the past 2,000 years averages about 0.0016 second per day per century, which amounts to an accumulated time difference of $29T^2$ seconds for a mean solar clock considered as having zero rate at 1900 A.D., where T is the number of centuries from 1900.[13] At present the deceleration is estimated at approximately 0.001 second/day/century, which amounts to an average apparent increase in frequency of about 1 part in 10^8 per century for an absolutely constant oscillator if measured in terms of the earth's rate of rotation.

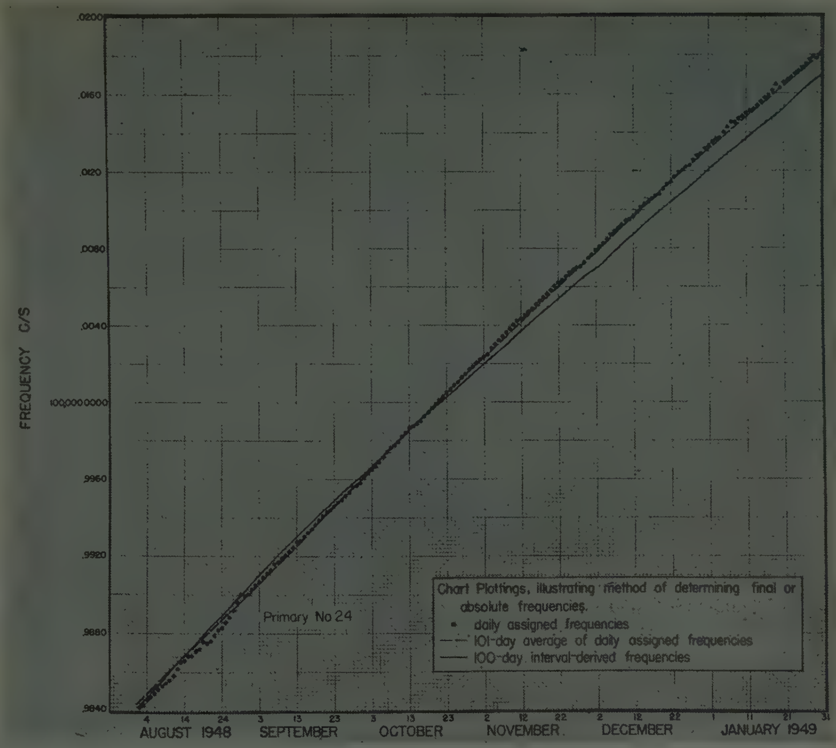


Fig. 7—Chart plottings illustrating graphical method of determining final or absolute frequencies.

The GT wire-supported crystals of the type widely used at present have initial drift rates of 2 to 4 parts in 10^8 per day for unaged units, and 1 to 2 parts in 10^9 per day for units a year old. The drift varies approximately inversely with time for several years, and may thus be predicted and allowances made accordingly.

In the primary standard of frequency of the National Bureau of Standards, three well-stabilized oscillators are evaluated over 100-day periods in terms of Naval Observatory time corrections. This is done as shown in Fig. 7. The dotted points for each day represent the departures from the average curve in terms of the entire group of selected standards as explained previously. The dashed curve represents the average of 101 daily values including 50 before and 50 after the day plotted. These are computed each 10 days and connected by a smooth curve. The solid curve represents the 100-day interval-derived average frequency computed for each day. The final corrected frequency, for this standard only, is then the daily plotted value of the interval derived (solid) curve plus the algebraic departure of the daily assigned (dotted point) value from the daily average (dashed) curve. A displacement of more than 1 part in 10^8 between these two curves indicates that the daily drift rates should be reassessed and the dotted points replotted to bring the two curves more nearly into agreement.

For a 100-day averaging period the daily assigned curves must thus be extrapolated or guided on the basis of past performance for about 60 days in advance of the established average curves. For this reason, a temporary 20-day interval-derived curve is computed as an aid in systematically determining the daily assigned values. Users of the WWV time signal transmissions in the field for average frequency determinations thus have the advantage of the long averaging period applied to the WWV frequency determinations. They may, therefore, approach the full limit of accuracy (1 part in 50 million) by using these signals over intervals of 6 to 10 days to compute average frequencies without knowledge of the time or extent of WWV frequency adjustments. If these adjustment data are known, accuracies several times this order are generally possible.

The corrected daily frequency for the reference oscillator is computed in terms of each of these three oscillators so evaluated by adding algebraically the respective daily beat differences to their corrected values. The mean of these computed frequencies is then taken as the final or "absolute" value of the reference for each day. Should one of the three values disagree excessively with the other two for explainable reasons, only two values are averaged. The values so computed generally agree within 1 or 2 parts in 10^9, which represents the residual error in the graphical method and the unpredictable random deviation of the daily assigned sampled values from the daily mean frequencies. The daily "absolute" frequencies for each of the other standards are obtained by algebraically adding their daily beat differences

[11] H. Spencer Jones, "The measurement of time," *Endeavour*, vol. 4, pp. 123–130; October, 1945.
[13] G. M. Clemence, "On the system of astronomical constants," *Astronomical Jour.*, vol. 53, pp. 169–179; May, 1948.

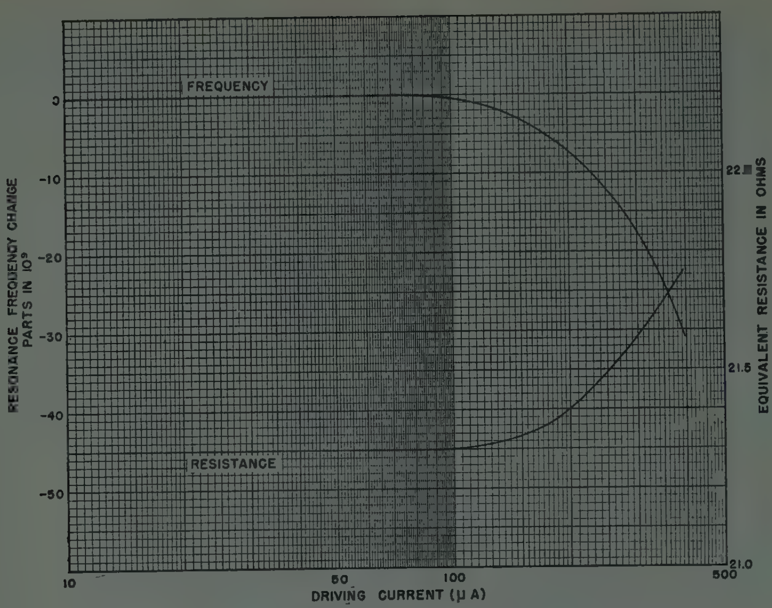


Fig. 8—Graphs showing variations of resonance frequency and series resistance at resonance with changes in driving current for a 100-kc GT-cut quartz-crystal unit.

to the daily "absolute" frequency of the reference standard.

No great difficulty has been experienced in predicting the "absolute" value of frequency for the primary standard in terms of the earth's mean rate of rotation over 100-day intervals to 1 part in 10^8. The mean solar second, which is by definition the physical standard of time, may be expected to change occasionally by as much as ± 4 parts in 10^8 in addition to a gradual average change of about 1 part in 10^8 per century. Frequency which is defined in terms of the mean solar second will, of course, deviate from a constant value by a corresponding amount.

Prospective Improvements in Quartz-Crystal Frequency Standards

Quartz-crystal units, measured as resonators in a balanced bridge network, have been compared with the primary standard of frequency with a precision of 1 part in 10^9. In this manner, crystal units most suitable for use in future standard oscillators are currently selected and studied. The use of crystal-unit resonators as primary frequency reference standards has been tried in this manner to some extent and is being further investigated. This method of use eliminates the variations arising from aging or detuning of tubes and circuit components. It has been found that the crystals generally drifted about as rapidly in a nonoscillating condition as they do in continuously oscillating frequency standards. Fig. 8 shows the results of bridge measurements on a typical GT-cut, wire-mounted crystal-unit for variation in excitation current. It may be noted that the frequency decreases and the resistance increases slightly with increases in excitation beyond about 100 microamperes through the crystal-unit at resonance. All crystal oscillator standards operating at present at The National Bureau of Standards are working at amplitudes of 500 microamperes or more, which is slightly beyond the maximum shown on the graph. Attempts to operate with appreciably lower amplitudes, by adjusting the bridge arms to closer limits, have resulted in increased short-time instability although the curves indicate greater stability should result. Extreme care in shielding, decoupling, and attention to the reduction of tube and circuit noise may be necessary to gain improvement in this manner.

Ability to operate the amplifier-bridge loop at a higher stable amplification ratio may be made possible by supplying the thermistor element for amplitude stabilization with direct current obtained from rectified and filtered oscillator output. This may also allow operation of the crystal unit at nearer its natural Q value. The present standards have crystal units with Q values of about 200,000 and operating Q values of about one-half this amount. This gives a bandwidth between half-power points of about 1 cps at 100 kc. The short-period stability of about 1 part in 10^{10} obtained with several of the standard oscillators means that they hold the frequency constant to about 1/100,000 of the half-power bandwidth. Current development work on improving crystal-units shows promise of greatly reducing the initial aging or drifting in frequency and of obtaining an operating Q factor of one million or greater.

The use of direct polarizing voltage across the crystal electrodes as a means of fine frequency adjustment may be advantageous in certain cases. A change in frequency of about 1 part in 10^8 per volt occurs for the present type 100-kc units. This change in frequency is very linear over ranges of several hundred volts of either polarity without apparent loss of Q. It offers an extremely sensitive control method for interpolation or servo applications without the problems that arise from backlash and wear.

Improved temperature control of the crystal inclosure over long periods is needed. A temperature-control method using dual resistance-bridge thermostats is planned for use in several new standard oscillators. The possibility of using magnetic amplifiers in this application to eliminate tube failures is being investigated.

Absolute Frequency and Time Standards

It is now generally accepted that the earth's mean rotation period is neither absolutely constant nor is it predictable with desirable accuracy over short or long periods. Its long-time period is increasing by an amount sufficient to be slightly disturbing to both physicists and astronomers in this era of high precision measurements.

Astronomers are using what is termed Newtonian time.[13] This time is consistent with Newton's laws of motion (with slight modifications for relativity) when applied to the movement of astronomical bodies. So far as has been determined, intervals in Newtonian time are invariable.

The Newtonian second for astronomical purposes has been defined as equal to a mean solar second at 1900 A.D. Astronomical events in the distant past are computable in Newtonian time by applying the proper corrections, based on the earth's known variable rate through the period considered. Future corrections, while not accurately predictable, can be observed and adopted as time progresses. The sidereal year (average period of the earth's revolution around the sun) is believed by astronomers to be a better unit of time. With present techniques it is difficult to determine the period of a single sidereal year to the desired accuracy. It has been estimated that in an interval of 100 years a mean value good to 1 part in 10^9 could be established. To subdivide this long interval into useful physical time units imposes extremely stringent requirements on a standard clock.

Clocks have continued to be improved since the discovery of the escapement mechanism about 1360 A.D., reaching their high degree of dependability in the present-day precision quartz clocks which permit a predictable daily constancy of considerably better than 0.001 second per day. They have an attainable precision of about 10 times that of the best mechanical clocks and are the most precise timekeepers now available, exceeding the constancy of the determination of the earth's rate for periods up to several months. The quartz clock, however, must be set in terms of the earth's mean rate, as it in no way constitutes an absolute standard in itself.

In recent years, aided by the rapid development of microwave techniques, considerable attention has been directed toward use of atomic resonance effects at microwave frequencies.[13]

The first atomic clock was built at the National Bureau of Standards in 1948.[14] This clock makes use of the sharp absorption line of ammonia gas at 23,870.1 Mc to maintain a 100-kc crystal oscillator at constant frequency. This is done by means of an electronic servosystem consisting of frequency

[13] W. D. Hershberger and L. E. Norton, "Frequency stabilization with microwave spectral lines," *RCA Rev.*, vol. 9, pp. 38–49; March, 1948.
[14] "The atomic clock," *NBS Tech. News Bull.*, vol. 33, pp. 17–24; February, 1949.

multipliers, auxiliary frequency-modulated search oscillator, and pulse discriminator circuits. The 100 kc is then divided down to audio frequencies in conventional manner and used to operate a synchronous-motor clock as a time standard. A constancy in frequency of 5 parts in 10^8 has been obtained for periods of several days with this experimental clock when compared with WWV frequency standards. Improvement of this type of standard and the development of more constant types of atomic resonance controlled oscillators are to be expected.

It is shown in the references cited that the sharpness of resonance within individual oscillating molecules is extremely great. Because of collisions of molecules with each other and with the gas cell walls, and the Doppler broadening attributable to natural thermal agitation, the practical working Q of an ammonia gas absorption line ranges between 50,000 and 500,000. This compares favorably with the Q of quartz crystals used in frequency standards which ranges from 100,000 to 1,000,000. It is thus reasonable to hope that a constancy of 1 part in 10^8 to 1 part in 10^9 may be obtained by proper refinements in circuitry and technique. Whether or not this degree of constancy of absolute value can be maintained without or even with precise temperature and pressure regulation remains to be investigated.

Civil time will, no doubt, continue to be defined in terms of the mean solar second, as would ordinary frequency designations. After about 2,000 years, if the earth continues to slow down at its present rate, the mean solar second would be about 1 part in 3 million longer than at present and the accumulated time difference between mean solar time and Newtonian time would amount to about three hours.

Conclusions

Frequency and time standards, using high-precision quartz crystals, are now available which are capable of supplying frequencies and time intervals constant to considerably better than 1 part in 10^8 per day. In order to achieve an accuracy approaching this order, these standards must be frequently checked in terms of standard frequency or time broadcasts. An accuracy of 1 part in 10^8 represents about the limit obtainable in terms of the earth's mean rate of rotation over a 100-day period. Longer periods of averaging can not be expected to give greatly improved accuracy, because of the possibility of slight oscillator frequency deviations and uncertainties in the uniformity of the determinations of the earth's mean rate.

By using ordinary zero beating methods, a remote frequency standard may be adjusted within 1 part in 10^7 to WWV's received frequency. Special offset techniques permit this setting error to be reduced to less than 1 part in 10^8. Changes in the radio propagation medium may cause the received frequency to differ from that transmitted by as much as several parts in 10^7. By averaging a number of determinations made when noon or midnight prevails about halfway between transmitter and receiver, long-distance frequency comparisons can generally be made with a precision of better than 1 part in 10^8.

The intercomparison of two remote oscillators, constant to 1 or 2 parts in 10^9 per day, by means of transmitted time pulses from one or both of the standards, is possible to a precision of a few parts in 10^9 through comparisons of average frequencies over periods of 6 or more days.

The development of atomic or molecular-resonance standards of high constancy and absolute accuracy may greatly simplify the maintenance of precise frequency and time standards. The practical realization of such standards, which seems reasonably probable in the near future, will eliminate the necessity of making highly precise physical measurements in terms of the earth's variable rate of rotation. Such a standard would supply a means of studying more precisely the motions of the earth and other astronomical bodies.

Considerable work is being done to improve the constancy of quartz-crystal frequency standards, especially with regard to aging or frequency drift and improved temperature control methods.

It is probable that frequency and time standards, which have improved by a factor of ten or more per decade in the last thirty years, will continue to reach new orders of accuracy and constancy. However, these improvements in accuracy will probably be referred to a new kind of standard, rather than to the mean solar second.

Acknowledgments

The author wishes to acknowledge his indebtedness for the valuable advice and assistance of W. D. George, Chief of the High-Frequency Standards Section of the National Bureau of Standards, and the many suggestions of V. E. Heaton and E. L. Hall of the same organization.

CORRECTION

It has been brought to the attention of the editors that the name of Han Chang, co-author of the paper, "The Reactance-Tube Oscillator," which appeared on pages 1330–1332 of the November, 1949, issue of the Proceedings of the I.R.E. was misspelled in his biography on page 1345 of that issue. The editors regret this inadvertent error.

The Speed of Radio Waves and Its Importance in Some Applications*

R. L. SMITH-ROSE†, FELLOW, IRE

Summary—**This paper comprises a review of the present state of knowledge of the speed of transmission of radio waves under the practical conditions of certain applications in which such knowledge is important. It is shown first that, for radio waves in a vacuum, their speed of transmission is equal to the velocity of light (299,775 km/s), to within the limits of experimental error. When waves of frequencies in the neighborhood of 100 kc/s are propagated at a height of a fraction of a wavelength above the earth's surface, their speed is reduced by an amount dependent upon the electrical conductivity of the earth. For overland transmission, the speed is about 299,250 km/s. For higher frequencies propagated at a height of several wavelengths, the speed of the waves is determined by the refractive index of the air, rather than by the properties of the ground. Since the refractive index decreases with the height of transmission, so does the speed of the waves increase toward the velocity of light. For example, centimeter waves propagated at heights of a few hundred feet have been observed to travel at a speed of about 299,690 km/s. When the waves are transmitted between ground and aircraft flying at a height of 30,000 feet (9,800 meters) this speed is increased to about 299,750 km/s.**

I. Introduction

THE DEVELOPMENT of various radio applications in the past few years, such as the exploration of the ionosphere, radar, and various navigational aids, has given rise to an independent need for an accurate knowledge of the velocity of radio, as distinct from light, waves propagated under various conditions. It is the purpose of the present paper to give a brief review of recent investigations in the subject and the resulting state of our knowledge.

II. The Velocity of Light and Radio Waves in a Vacuum

(a) *The Velocity of Light*

Several independent reviews[1–4] of the results of direct measurements of the velocity of light have been made in recent years, some of these having been directed toward establishing the precise value of the velocity in vacuum, which is regarded as one of the most useful physical constants. It has been known and appreciated for a long time that the speed of light waves traveling through air or any other medium would vary with the permittivity or dielectric constant of the medium in accordance with Maxwell's original conception. But physicists have been a little perturbed by the suggestions made from time to time that the value of the velocity was varying slowly from year to year. The available evidence on this point was carefully considered by Birge[1] in his review of the General Physical Constants in 1941, and he concluded that the value of

$$299,776 \pm 4 \text{ km/s}$$

could be ascribed to the velocity of light in a vacuum, and that there was no satisfactory indication of any secular change in this value.

Another and rather more detailed review of the position was published in 1944 by Dorsey,[3] who with E. B. Rosa at the Bureau of Standards over forty years ago had determined the value of the velocity from calculations and measurements made on the capacitance of a capacitor of suitable shape. He confirms that the velocity of light in a vacuum is invariable but concludes that the best value for it is

$$299,773 \pm 10 \text{ km/s}.$$

From such considerations of the now classical experimental measurements of the speed with which light waves travel in a vacuum, we may conclude that this velocity is known to an accuracy of a few parts in 10^6, and that its value is less than the conveniently assumed figure of 300,000 km/s by about 750 parts in a million.

In the remainder of this paper, the value of the velocity of light waves in vacuum will be taken to be the constant:

$$c_0 = 299,775 \text{ km/s}$$
$$= 186,272 \text{ miles per second.}$$

(b) *Measurements on Radio Waves in a Vacuum*

Over twenty years ago, Mercier[5] described measurements of the velocity of electromagnetic waves guided by parallel wires in air. By making corrections for the assumed effects of the wires and atmosphere, a value for the resulting velocity of free waves in air was deduced.

Quite recently, Essen and Gordon-Smith[6] have considerably improved on this technique at the National Physical Laboratory in England, in the course of the development of cavity resonators for centimeter wave-

* Decimal classification: R111.1×R500. Original manuscript received by the Institute, April 15, 1949; revised manuscript received, August 5, 1949. Presented, 1949 IRE National Convention, New York, N. Y., March 9, 1949.

† Department of Scientific and Industrial Research, Radio Research Station, Slough, England.

[1] R. T. Birge, "The general physical constants," Reports on Progress in Physics, vol. 8, p. 92; 1941.

[2] R. L. Smith-Rose, "The speed of travel of wireless waves," *Jour. IEE*, part 1, vol. 90, pp. 31–83, January, 1943.

[3] N. E. Dorsey, "The velocity of light," *Trans. Amer. Phil. Soc.*, part 1, vol. 34, pp. 1–110; 1934.

[4] J. Warner, "The velocity of electromagentic waves," *Australian Jour. Sci.*, vol. 10, pp. 73–76; December, 1947.

[5] J. Mercier, "On harmonic and multiple synchronization," *Jour. Phys. and Radium*, vol. 5, pp. 173–179; 1924.

[6] L. Essen and A. C. Gordon-Smith, "The velocity of propagation of electromagnetic waves derived from the resonant frequencies of a cavity resonator," *Proc. Roy. Soc.*, vol. 194 A, pp. 348–361; 1948.

lengths. A cylindrical resonator was turned out of solid copper, and its internal dimensions were measured very carefully. From such measurements, the resonant frequency of the cavity can be calculated from the formulas for the propagation of waves in a cylindrical waveguide, and taking account of the well-known skin effect at the relevant radio frequency. This calculation involves a knowledge of the velocity of propagation of the radio waves inside this short, closed waveguide system. By measuring the resonant frequencies of the cavity for various modes, the above investigators were thus able to determine this velocity. The measurements were made with the resonator in a vacuum and led to a result of

$$299{,}792 \pm 9 \text{ km/s}$$

for the velocity of radio waves at a frequency in the region of 3,000 Mc/s.

This result obtained for a frequency of 3×10^9 cps (wavelength 10 cm) is about 17 km/s greater than that quoted above for light waves having a frequency of about 5×10^{14} cps (wavelength 6×10^{-5} cm). This difference is greater than the probable inaccuracy of either observation claimed by the workers at the two respective frequencies, so that it must remain for further investigation to resolve the apparent discrepancy in the determination of the physical constant denoted by c_0.

III. The Speed of Radio Waves under Conditions of Practical Application

The experiments just described refer to measurements made under laboratory conditions where, for standardizing purposes, the highest precision is required. Although it was long ago realized that, in the practical use of radio waves for communication and other purposes, the speed of travel of the waves would be affected by the transmission conditions over the ground and through the atmosphere, the need for a precise investigation of this subject has become increasingly apparent during the past decade.

In the first place, it is clear that the whole field of radar technique and its many applications is based upon a measurement of the time of transit of radio waves from a sender to a target and back to a receiver; and the distance of the target, which is one of the main factors required, can only be deduced with the aid of a knowledge of the speed of the waves.

Secondly, one of the most important results of the exploitation of radio-wave technique during the war years has been the development of new and improved aids to air and marine navigation. Some of these, such as the now well-known Gee and Loran methods, comprise a means of fixing the position of a radio-receiving point by determining the difference in time of arrival of pulse-modulated signals emitted in precise synchronism by two or more transmitting stations. Other systems such as Decca use continuous waves, and measure the same time difference of the arriving signals in terms of the phase difference of the carrier waves arriving at the receiver from two or more sending stations. In any of these cases, the time difference is measured by reference to a standard frequency of oscillation provided either by the transmitting station or incorporated within the receiving equipment. For the translation of time into distance, however, it is necessary to have a precise knowledge of the speed with which the waves travel between the transmitting stations and the receivers at which the measurements are made. In some applications, it will also be necessary to distinguish between the phase and group velocities of the waves. In the case of transmissions to aircraft, the velocity required may be that appropriate to the travel of waves through the air free from the effects of any obstacles on the ground, but subject to the effects, if any, of variations in atmospheric conditions such as density and moisture content. When the position finding is carried out on board ship, the waves will usually have traveled along the surface of the earth, perhaps partly over land and partly over sea, and it is necessary to know the speed with which the waves have traveled over the particular terrain forming the path between the transmitting and receiving points. In assessing the accuracy of such methods of using radio waves for position-fixing and navigational purposes, it is important to appreciate the limitations imposed by our knowledge of the speed of propagation of the radio waves, and it is already realized that there is need for further investigation on this subject if full advantage is to be taken of these important developments.

IV. The Speed of Radio Waves Transmitted over the Earth's Surface

In a review[2] of this subject made in 1942, the author drew attention to the results obtained some years earlier in the course of an investigation on the velocity of medium radio waves carried out in the USSR under L. Mandelstam and N. Papalexi. From the observations then available, it was concluded that limiting values of 299,000 and 299,500 km/s could be ascribed to the speed of such waves for transmission over a clear air path or over sea or fresh water. In the case of transmission over land, a somewhat lower value of 295,000 km/s was obtained, although this would appear to be subject to some uncertainty in determining the true path of transmission, owing to various intervening obstacles. In later work, the Russian authors have described their investigations into the nature of the electromagnetic field near an antenna, and the effect of distance on the speed of transmission. As pointed out by Ratcliffe,[7,8] the results of this work show that the velocity of the ground wave is less than the velocity in free space, provided we are not too far from the transmitter, and equal

[7] J. A. Ratcliffe, "The velocity of radio waves," *Proc. Union Radio Sci. Internationale* (Paris), vol. 6, p. 107; 1946.

[8] J. A. Ratcliffe, "A source of error in radio navigation systems which depend upon the velocity of a 'ground wave'," Proc. I.R.E., vol. 35, p. 938; September, 1947.

to that in free space at greater distances. These conclusions are consistent with the earlier calculations of Norton,[9,10] who has also pointed out the importance of knowing the ground-wave transmission conditions in the modern application of radio navigation systems. The work of the USSR investigators has also been supplemented by Eckersley,[11] in an investigation of coastal refraction effects. He claims that the theoretical explanation of this phenomenon is now satisfactory, and that what experimental evidence is available supports the conclusion that the refraction is only significant over a limited frequency range of from about 200 to 20,000 kc/s (wavelengths 15 to 1,500 meters), and that the effect disappears at distances sufficiently far from the coastal boundary. More recently, Millington[12,13] has described an investigation in this field.

While much of the work referred to above is of a quantitative nature, and indicates by how much the velocity of radio waves transmitted over land and sea is less than the value for free-space conditions, it is all of a rather low order of accuracy for the modern requirements of, say, a precise navigational system. It was in the course of the experimental use and calibration of such a navigation system that a method of finding the effective velocity of propagation of radio waves at frequencies in the region of 100 kc/s was developed by the British Admiralty Signal Establishment in 1945, and the results obtained by this method have been described by Mendoza.[14]

The measurements were made with the aid of a chain of three Decca transmitters operating on frequencies of 85 kc/s, 113.33 kc/s ($4/3\times 85$), and 127.5 kc/s ($3/2\times 85$) respectively; and observations were made in an airplane of the complete phase change experienced in flying round the system at a height of 1,000 feet (330 meters) which is about one-tenth of a wavelength above the ground. The measurements were made mostly at distances between 15 and 100 km along the extensions of the base lines connecting pairs of stations, but no significant change of readings with distance could be detected, nor were there any signs of dispersion within the limits of experimental error which were estimated at 1.4 parts in 1,000.

The mean value for the velocity of the radio waves determined under the above conditions was found to be

$$299{,}250 \pm 40 \text{ km/s}.$$

This is in good agreement with the results obtained several years earlier by the USSR investigators, although the precision of their measurements was much less. Also, as Ratcliffe pointed out in the discussion on Mendoza's paper, the above value of velocity would correspond from Norton's calculations to propagation over land having a resistivity of 3×10^3 ohm-cm (conductivity 10^8 electrostatic units). This would seem to be appropriate to the type of land over which the measurements were made.

Experience obtained in the use of another chain of Decca stations operated over land of lower conductivity (about 0.5×10^7 electrostatic units) has suggested that the velocity of the ground waves in this case was about 298,100 km/s; but these measurements are not of the same precision as those quoted above. When the propagation is over sea, the speed of the waves approaches that for free air transmission at the lower frequencies under consideration.

As is not uncommon in investigations of the type just described, the accuracy of the method is limited by the very observations which the navigational system is designed to avoid, namely, the determination of the position of the aircraft at the time of making the radio observations. This difficulty would naturally be increased in attempts made to repeat the measurements entirely over sea; although it is clearly desirable that, when practicable, the velocity measurements should also be made for conditions of transmission over sea. This, in turn, will lead to a more detailed study of transmission over a mixed land and sea path, which is a matter of great interest to those concerned with ground-wave radio propagation.[12,13]

V. The Speed of Very Short Radio Waves Transmitted through the Lower Atmosphere

It is well known that when electromagnetic waves travel through a medium of dielectric constant ϵ, their speed is inversely proportional to the refractive index n of the medium where $n=\sqrt{\epsilon}$. Thus when the waves, whether corresponding to light or radio frequencies, are transmitted through the air their velocity will be reduced to $1/n$ of the value in a vacuum. If we express n as equal to $1+\alpha$, then for air at the earth's surface in temperate latitudes, the value of α is of the order of 300×10^{-6}, but it may vary with temperature, pressure, and humidity from values of 400×10^{-6} or so down to negligibly small values, as the pressure decreases to zero. Thus we should expect the velocity of light in vacuum as given in Section II to be reduced to 299,675 km/s for transmission through air having a refractive index of about $1+350\times 10^{-6}$ which corresponds approximately to a pressure of 760 mm, a temperature of 18°C and a relative humidity of 70 per cent.

Now the development of radar during the war led to

[9] K. A. Norton, "The calculation of ground-wave field intensity over a finitely conducting spherical earth," Proc. I.R.E., vol. 29, pp. 623–640; December, 1941.

[10] K. A. Norton, "A new source of systematic error in radio navigation systems requiring the measurement of the relative phases of the propagated waves," Proc. I.R.E., vol. 35, p. 284; March, 1947.

[11] T. L. Eckersley, "Coastal refraction," *Atti del Congresso Internazionale per il Cinquantenario della scoperta Marconiana della Radio*, p. 97, 1947.

[12] G. Millington, "Ground wave propagation across a land-sea boundary," *Nature*, vol. 163, p. 128; January, 1949.

[13] G. Millington, "Ground wave propagation over an inhomogeneous smooth earth," *Jour. IEE*, part III, vol. 96, pp. 53–64; January, 1949.

[14] E. B. Mendoza, "A method of determining the velocity of radio waves over land on frequencies near 100 kc/s," *Jour. IEE*, part III, vol. 94, pp. 396–399; November, 1947.

the use of some very precise methods of air navigation, and these have provided means of measuring the actual velocity of transmission through air for meter and centimeter waves. The results of such measurements, which have recently been published in England, have provided an experimental confirmation of this effect of atmospheric conditions on the speed of the waves.

The first series of measurements have been described by Smith, Franklin, and Whiting,[15] and were made with the aid of radio stations operating on the air navigation system G-H., at frequencies between 22 and 60 Mc/s. The technique adopted was to emit pulses of waves from an "interrogator" station, and these were received and re-transmitted from two "responder" stations at distances of 125 and 140 km respectively. The total time taken for the waves to go and return along each path was measured with the standard Gee equipment, steps being taken to determine and allow for the time of transit of the pulses through the equipment and feeders. Four experiments were carried out on separate days, and the observations indicated that the over-all accuracy was better than 2 parts in 10^4. Although the paths of the waves in the case of tests with one responder station were entirely over sea, while, in the other case, the path was partially over hilly land, there was no significant difference in the results obtained. It is considered, therefore, that the measurements apply to the velocity of meter waves through air at or near sea level, and the mean value obtained was

$$299{,}695 \pm 5 \text{ km/s}.$$

This is within the range of values expected from considerations of the atmospheric conditions referred to above.

The next series of measurements, described by Jones,[16] were carried out in a similar manner but with the range-measuring system known as Oboe. In this case the experiments, which were conducted in two parts, were made with pulse emissions on a wave frequency of about 3,300 Mc/s (wavelength 9 cm). The first series of measurements were made over two clear sea paths, 50 and 67.5 km long, using one control and two responder stations on the ground. As before, steps were taken to ensure that any time delays in the passage of the pulse signals through the equipment was allowed for, and it was found that there was no significant difference in the velocity measured over the two paths. Thus the mean value of several measurements was found to be

$$299{,}687 \pm 25 \text{ km/s}$$

which is in remarkably good agreement with the result obtained on meter waves under the same conditions of transmission over sea at ground level.

TABLE I

Height of Aircraft	*Mean Velocity*
10,000 feet	299,713 km/s
20,000 feet	299,733 km/s
30,000 feet	299,750 km/s

TABLE II

Summary of Measurements of the Speed of Radio Waves

Authority	*Waves*		*Conditions of Transmission*	*Mean Value in km/s*	*Accuracy of Measurement*	
	Type	*Mean Frequency cps*			*km/s*	*Parts in 10^4*
Various	Light	5×10^{14}	In vacuum	299,775	± 10	0.3
	Radio	*Mc/s*				
L. Essen and A. C. Gordon-Smith	Continuous	3,000	In vacuum	299,792	± 9	0.3
L. Mandelstam and N. Paplexi	Continuous	1	Ground to ground	299,250	±300	10
E. B. Mendoza	Continuous	0.1	Ground to ground over land	299,250	± 40	1.3
R. A. Smith, E. Franklin, and F. B. Whiting	Pulsed	40	Ground to ground over sea	299,695	± 50	1.7
F. E. Jones	Pulsed	3,300	Ground to ground over sea	299,687	± 25	0.8
F. E. Jones	Pulsed	3,300	Ground to 3,300 meters Ground to 6,500 meters Ground to 9,800 meters	299,713 299,733 299,750	—	—

[15] R. A. Smith, E. Franklin, and F. B. Whiting, "Accurate measurement of the group velocity of radio waves in the atmosphere, using radar technique," *Jour. IEE*, part III, vol. 94, pp. 391–396; November, 1947.

[16] F. E. Jones, "The measurement of the velocity of propagation of centimeter radio waves as a function of height above the earth"; Part I, "Ground-level measurement of the velocity of propagation over a sea path," *Jour. IEE*, part III, vol. 94, pp. 399–402; November, 1947. Part II, "The measurement of the velocity of propagation over a path between ground and aircraft at 10,000, 20,000, and 30,000 feet," *Jour. IEE*, part III, vol. 96, pp. 447–452; September, 1949.

For the second series of experiments, the measurements were made between two ground stations and an aircraft flying at various heights. Although it was possible to make only a limited number of flights, the values obtained for the mean velocity of transmission between ground and aircraft were as listed in Table I.

These results confirm very well, indeed, the expectation that, as the height of the atmosphere through which the transmission takes place is increased, so does the speed of the waves tend toward the value at which they travel in a vacuum.

VI. Summary

The present state of our knowledge of the speed of radio waves may be summarized by the values given in Table II, in which the best known value for the velocity of light in a vacuum is included.

The Application of Thermistors to Control Networks*

J. H. BOLLMAN†, SENIOR MEMBER, IRE, AND J. G. KREER†, SENIOR MEMBER, IRE

Summary—**In connection with the application of thermistors to regulating and indicating systems, there have been derived several relations between current, voltage, resistance, and power which determine the electrical behavior of the thermistor from its various thermal and physical constants. The complete differential equation describing the time behavior of a directly heated thermistor has been developed in a form which may be solved by methods appropriate to the problem.**

I. General

One of the principal features of the newer cable carrier telephone systems[1,2] is the extensive use of a new circuit element, the "thermistor," for various circuit functions requiring a remotely controlled variable or nonlinear resistance. The use of this new element in the gain regulating system permits a considerable saving in space and cost together with an improvement in system gain stability.

The word "thermistor" is a contraction of the words thermal resistor, and designates a type of circuit element, the electrical resistance of which varies over a wide range with changes in temperature. In contrast with metals which have small positive temperature coefficients of resistance, thermistors are made from a class of materials known as semiconductors which have relatively large negative temperature coefficients, that is, the resistance decreases markedly as the temperature increases. A particularly valuable type of thermistor is made of a number of metal oxides sintered into a compact mass at high temperatures. The specific resistance at a given temperature and the temperature coefficient depend upon the relative proportions of the oxides and upon the heat treatment of the unit. A typical resistance-temperature characteristic of a thermistor material is shown in Fig. 1. It will be noted that the specific resistance is about 20,000 ohm-centimeters (resistance between opposite surfaces of a 1-centimeter cube) at 0°F. and about 20 ohm-centimeters at 400°F. or a range in resistance of 1,000 to 1 over this temperature range.

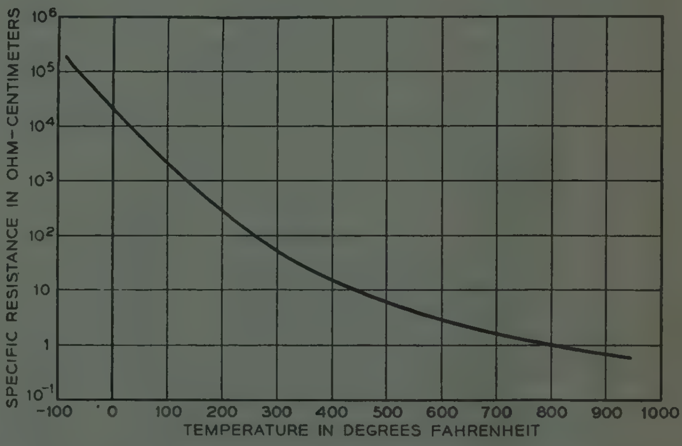


Fig. 1—Specific resistance versus temperature of the thermistor material used in some carrier telephone applications.

In the various regulating and indicating systems which have been developed, several types of thermistors are used. One is the disk-type thermistor in which a relatively large mass of thermistor material is compressed in the form of a disk and used as a resistance element which is not heated to any appreciable extent by the current passing through it. The resistance is therefore determined by the ambient temperature. Another type is that in which a small coil of wire is wound around an extremely small thermistor bead to form a heater which

* Decimal classification R282.12. Original manuscript received by the Institute, October 24, 1949.

† Bell Telephone Laboratories, Inc., Murray Hill, N. J.

[1] H. S. Black, F. A. Brooks, A. J. Wier, and I. G. Wilson, "An improved cable carrier system," *Trans. A.I.E.E.*, vol. 66, pp. 741–746; 1947.

[2] J. H. Bollman, "A pilot channel regulator for the K-1 carrier system," *Bell Lab. Rec.*, vol. 20, pp. 258–262; June 10, 1942.

serves to vary the temperature of the bead and thereby change its resistance. This is known as the indirectly heated type. The third type is that in which a small thermistor bead is heated by the current through itself. This is known as the directly heated type. The latter two types are frequently mounted in vacuum and otherwise treated to decrease the heat losses and increase the sensitivity.[3]

In the normal operating range, the life of a thermistor element is independent of the current flowing through it. Laboratory tests show that the resistance may change slightly (2 or 3 per cent) after several years of normal operation. It is important to note, however, that excessive currents will greatly shorten their life, and units may be permanently damaged if such currents are used.

We will now derive several relations between current, voltage, resistance, and power which have been found useful in circuit design.

II. Steady-State Behavior

It has been shown[3] that the resistance of a semiconductor such as is used in thermistor units is related to the temperature of the material by the relation

$$R = \alpha \exp\left(\frac{\beta}{T}\right), \tag{1}$$

where

α and β are constants determined by the geometry and composition of the semiconductor and

T is the absolute temperature.

To a close approximation this equation holds over any temperature region in which no change of state occurs in the semiconducting medium. In applying the relation to the determination of the thermistor characteristics, it is assumed that the temperature of the thermistor is uniform. This, of course, is not strictly true when appreciable temperature gradients are set up by power dissipation. The determination of these gradients and the assignment of boundary conditions would be virtually impossible because of the small size and irregular geometry of these thermistor elements. However, we may derive a power-resistance relation neglecting the temperature gradient in the thermistor and determine experimentally the corrections to be applied.

In addition to the resistance-temperature equation (1) it is necessary to know the relation between the power P dissipated in the thermistor and its consequent temperature rise. This is of the form

$$P = k_c(T - T_0) + k_R(T^4 - T_0^4), \tag{2}$$

where k_c is the thermal conductance from the thermistor in watts per degree Centigrade, k_R is the radiation coefficient in similar units, and T_0 is the ambient temperature and likewise the thermistor temperature with no power dissipation. It has been assumed and experimentally verified that, in the working range, the temperature rise is proportional to the power dissipated in the thermistor. This is equivalent to assuming either that the thermistor is sufficiently well shielded against radiation so that this power loss may be neglected, in which case the constant of proportionality will be independent of temperature, or alternatively that only the linear component of the radiated power need be considered, in which case the constant of proportionality may be written

$$K = \frac{1}{k_c + 4k_R T_0^3}, \tag{2a}$$

where K is the temperature rise for a given power dissipation.

Starting with these two relations it is shown in the Appendix (13) that the natural logarithm of the ratio of the unheated resistance R_0 to the heated resistance R, $\ln R_0/R$, is related to the power through the equation,

$$\ln\frac{R_0}{R} = \frac{P}{\dfrac{T_0}{\beta}P + \dfrac{T_0^2}{K\beta}} = \frac{P}{SP + P_0}, \tag{3}$$

in which S is written for T_0/β, a dimensionless constant, and P_0 is written for $T_0^2/K\beta$ (dimensionally a power). Hence if we divide the power dissipated in the bead by the natural logarithm of the resistance ratio and plot the result against the power dissipated, we should obtain a straight line provided that the assumptions discussed in the preceding paragraphs are satisfied.

Experimentally it has been found that this plot is almost linear, but that the slope S and the intercept P_0 are not T_0/β and $T_0^2/K\beta$, respectively, but more complicated expressions because of the approximations involved in (1) and (2) and, in some cases, the presence of appreciable temperature gradients. In all cases, however, the experimental data follow closely the form of (3), making this expression extremely useful to the circuit designer.

It is shown in the Appendix (15) by a comparatively simple algebraic transformation that the current-voltage characteristics of these elements may be represented as a single parameter family of curves relating dimensionless variables, one of which is proportional to the current and the other proportional to the voltage across the unit. This relation is:

$$\frac{v}{u} = \exp\left[\frac{-uv}{Suv + 1}\right] \tag{4}$$

where

$$u = I\sqrt{\frac{R_0}{P_0}}$$

and

$$v = \frac{E}{\sqrt{R_0 P_0}},$$

[3] J. A. Becker, C. B. Green, and G. L. Pearson, "Properties and uses of thermistors—thermally sensitive resistors," *Trans. AIEE*, vol. 65, pp. 711–725; November, 1946.

I being the current in the thermistor and E the voltage across it. By means of this relationship it is possible to calculate specific values of u and v, although it is not possible to solve the equation for one of the variables in terms of the other because of the transcendental nature of the relation. Fig. 2 shows several members of this one-parameter family of curves in which the logarithm of v is plotted against the logarithm of u. The reason for so plotting the curves is that it is then possible to obtain the actual E-I characteristic of any particular thermistor by merely translating the curve in such a way as to maintain the axes parallel to their original position. It thus becomes possible in a design application to plot the required E-I relation of the thermistor on similar graph paper and superpose the two curve sheets until the required curve corresponds to one of the curves on Fig. 2. From the values of u and v which correspond to one unit of voltage and one unit of current respectively, one can calculate the values of S, R_0, and P_0 required to produce the desired unit. Measured values from the E-I curve of a thermistor having $R_0=45{,}155$ ohms, $S=0.130$, and $P_0=0.000654$ watts are shown on the curve.

In the design of circuits such as gain regulators, expandors, oscillators, and modulators which employ thermistors, it is desirable to predict the behavior of the thermistor when connected in almost any type of resistance network.

The following relations will hold if the applied voltage is not alternating or if the frequency of alternation is high enough so that the thermistor does not change its temperature to any appreciable extent over the voltage cycle.

Perhaps the simplest case is a directly heated thermistor and a resistance R in series with a generator of zero internal impedance. In this case the current I through the thermistor and the external resistance is the same, and the terminal voltage E is the sum of the voltage drops. To obtain a dimensionless variable v' which will be proportional to the voltage across a thermistor and resistance in series, note that the terminal voltage $E=V+IR$ and let

$$v' = \frac{E}{\sqrt{R_0P_0}} = \frac{V}{\sqrt{R_0P_0}} + \frac{IR}{\sqrt{R_0P_0}},$$

then by substitution from (4) the definition of u and v,

$$v' = v + u\frac{R}{R_0}\cdot$$

This curve may be constructed by adding to the ordinates of the generalized thermistor characteristic a straight line defined by the equation

$$v' = u\frac{R}{R_0}\cdot$$

This line and some of the more significant of the combined curves are shown on Fig. 3. The equation for these curves is

$$\frac{v'}{u} - \frac{R}{R_0} = \exp\left[\frac{-v'u + u^2\dfrac{R}{R_0}}{1 + Sv'u - Su^2\dfrac{R}{R_0}}\right]. \tag{5}$$

A value of S equal to 0.1125 has been used for all calculations in this memorandum, since this is the value of S for the material used in most of the gain regulating thermistors.

The usefulness of the curves may be easily seen in the design of a simple voltage regulator consisting of a thermistor and an ohmic resistance in series designed to maintain a specified voltage across its terminals. Suppose that such a regulator is required to maintain a voltage of 3.0 volts ±2.5 per cent for a current range of 0.5 to 5.0 milliamperes. Examining the curve on Fig. 3 marked $R/R_0=0.020$, we find that the peak occurs at $v'=0.69$ and the trough at $v'=0.66$, a spread in the voltage factor of ±2.22 per cent. On this curve the value of

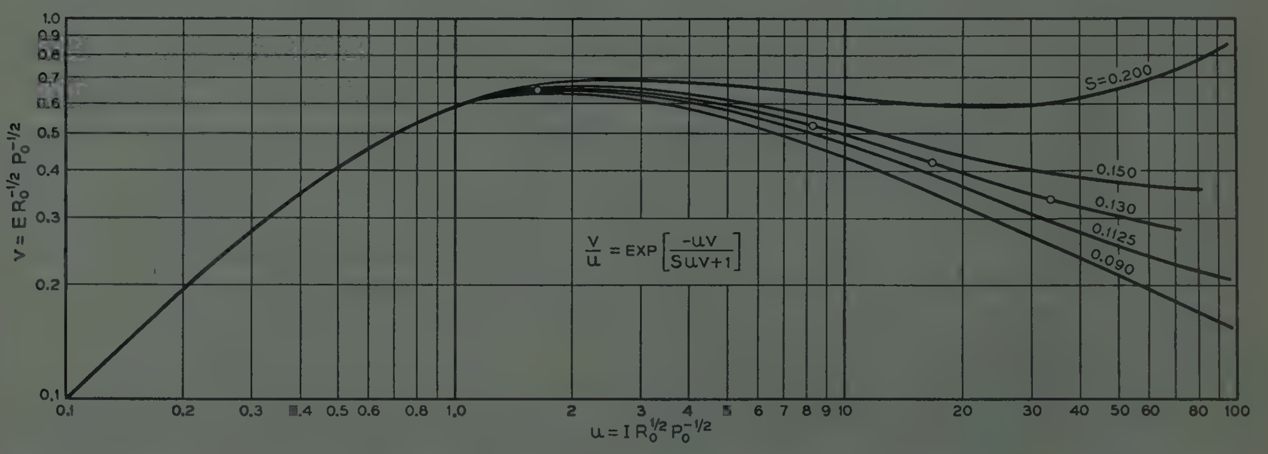


Fig. 2—Generalized thermistor voltage versus current characteristics. The open circles are measured values on a thermistor having $R_0=45{,}155$ ohms, $S=0.130$, and $P_0=0.000654$ watts.

v' stays within the ±2.5 per cent limits from $u=1.4$ to $u=16$, which is greater than the required 10 to 1 spread, consequently this value of $R/R_0=0.020$ may be used. The mean value of v' is then 0.675, which will correspond to 3 volts. The current factor may be evaluated by setting $u=1.5$ to correspond with a current of 0.0005 amp. (or $u=15$ to correspond with a current of 0.005 amp.). Using these relations to evaluate R_0 and P_0, the parameters of the thermistor will be $R_0=13{,}350$ ohms, $P_0=1.48$ milliwatts, $S=0.1125$. From the value $R/R_0=0.020$, a resistance of 267 ohms is required in series with the thermistor.

In Fig. 3 it may be seen that for values of R/R_0 less than 0.0297 the u versus v' curve and, therefore, the E versus I curve shows a definite maximum and minimum of voltage. At these points the incremental resistance in circuit must be zero, since $(dE/dI)=0$, and therefore the negative incremental resistance R_{AC} of the thermistor is just equal to the external resistance R. When $R/R_0=0.0297$, the u versus v' curve shows no maximum but only a point of inflection. This point then marks the maximum negative incremental resistance. In the Appendix (16), the relation between the resistance R_{AC} and the quantity $X=\ln R_0/R_b$ is obtained. This is plotted on Fig. 4, Curve I.

A curve (II) is also given by which the value of E may be determined when $R_{AC}=R$. By means of these curves it is possible to determine the accuracy of a voltage regulating network, since the points of maximum and minimum voltage mark the points of maximum departure from the desired output voltage.

It is sometimes desirable to determine the maximum negative incremental resistance which can be developed by a given thermistor. The equations which have been developed show that this maximum negative resistance

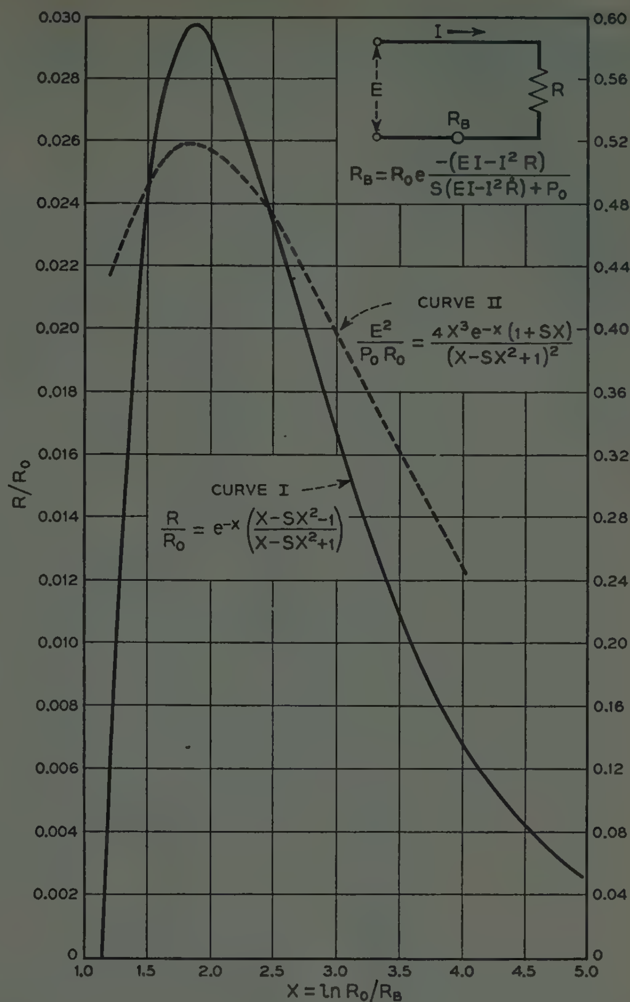


Fig. 4—Curves showing E^2/P_0R_0 and R/R_0 as functions of $\ln R_0/R_B$ when $(dE/dI)=0$ (E is at its minimum or maximum value).

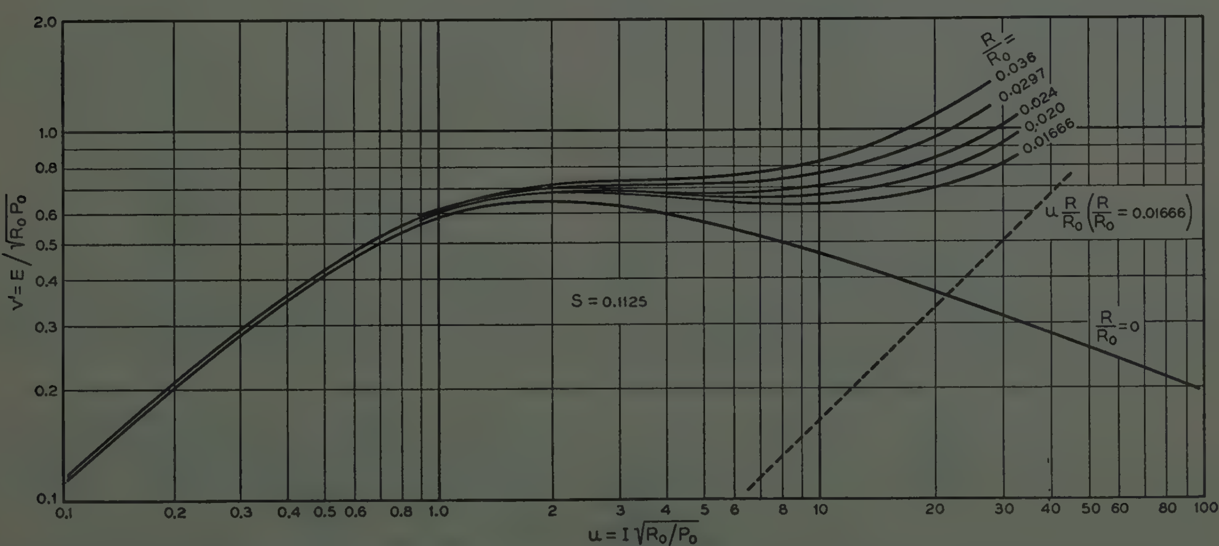


Fig. 3—Generalized thermistor voltage versus current characteristic modified to show the effect of adding series resistance.

will depend upon the value of S and R_0 only. Appendix (18) shows the relation between S and the value of x which gives the maximum negative resistance. This relation is plotted on Fig. 5. While it is inconvenient to

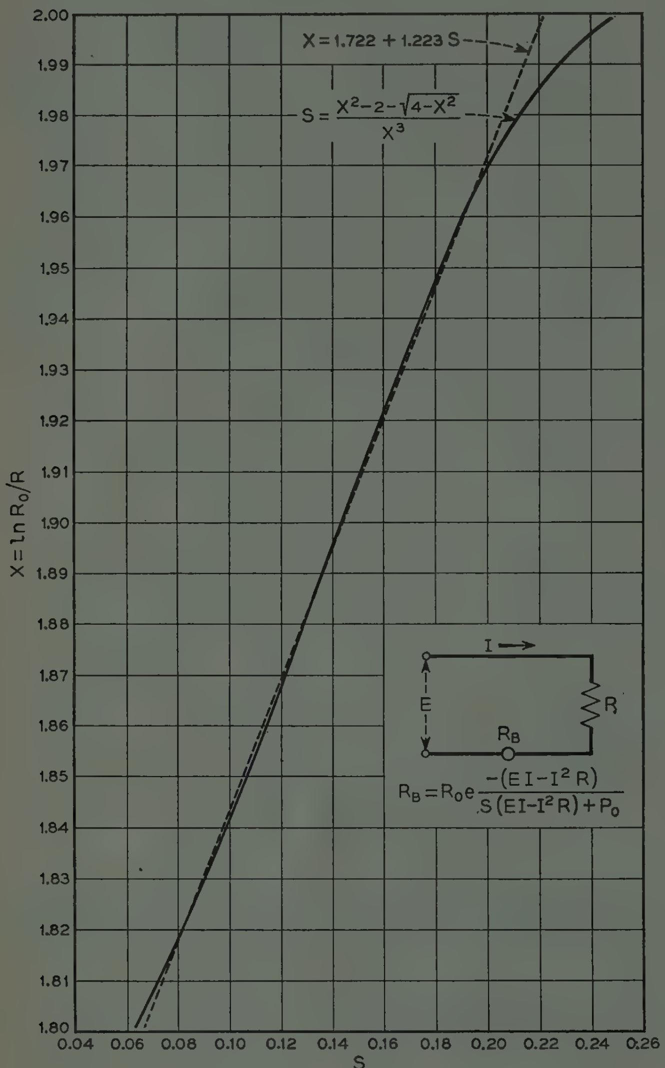


Fig. 5—Curves showing ln R_0/R as a function of S when the curve of E versus I shows no minimum or maximum, but only a point of inflection.

obtain x explicitly as a function of S, it may be seen that the straight line is a very close approximation to the actual curve, and this may be used if the accuracy is sufficient. To obtain the value of maximum negative incremental resistance, the value of S as determined by the composition of the thermistor bead is found on the horizontal axis of Fig. 5; finding the corresponding point on the curve, the value of x is determined. Using this value, the maximum negative resistance may be computed from the formula given on Fig. 4, or read from the curve on Fig. 6.

Another simple case is when the circuit configuration is such that a thermistor is in parallel with a resistance and working out of an infinite impedance generator. The voltages across the thermistor and resistance are the same or $E_b = E_R = E$ and

$$I = I_b + I_R = \frac{E}{R_b} + \frac{E}{R},$$

where the subscript b refers to the thermistor and R to the ohmic resistance.

By substitution

$$u_1 = u + v\frac{R_0}{R}.$$

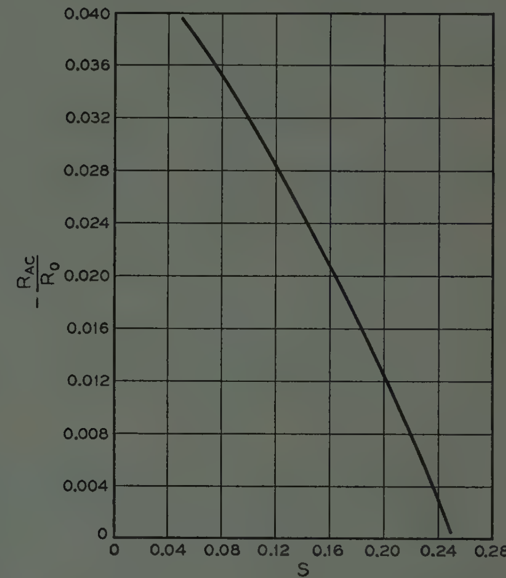


Fig. 6—The ratio of the maximum negative incremental resistance of a thermistor to its initial resistance as a function of S.

Hence this curve may be constructed by adding to the abscissa of the generalized thermistor characteristic a straight line defined by

$$u_1 = v\frac{R_0}{R}.$$

Having determined the characteristics of thermistors in series and parallel with single resistances, the characteristics of more complicated structures may be determined by well-known network design methods.

III. Dynamic Behavior

In the application of thermistors to certain circuits it is necessary to know how the thermistor resistance responds in time to changes in its input. This is particularly true in systems which employ a directly heated thermistor in the feedback circuit for regulation purposes.

To solve this problem, the differential equation will be set up which relates the resistance of a thermistor to driving force (the input current, voltage, or power) as a known function of time.

We will assume that the thermistor consists of a semiconducting bead having a thermal capacity, $m\sigma$ (mass times specific heat). This includes the thermal capacity

effects of leads, glass loading, radiation shields, and any other matter in intimate thermal contact with the bead.

We will also assume as in section II of this paper that when the bead is maintained at a temperature T the rate of energy loss through thermal leakance is proportional to the rise in temperature above some fixed ambient temperature which we will denote by T_0. Under these conditions at any instant the electrical power supplied to the thermistor equals the heat power loss through thermal leakance plus the rate at which thermal energy is being stored in the bead by the increase in temperature, i.e.,

$$m\sigma \frac{d(T - T_0)}{dt} + K(T - T_0) = W(\lambda). \qquad (6)$$

This is a linear differential equation with constant coefficients provided $W(\lambda)$ is independent of T. The solution can be written in the form:

$$T - T_0 = \left[c + \frac{1}{m\sigma}\int_0^t W(t) \exp\left(\frac{Kt}{m\sigma}\right) dt\right] \exp\left(\frac{-Kt}{m\sigma}\right), \qquad (7)$$

where c is a constant determined by boundary conditions.

Now the thermistor resistance is given by $R(t) = \alpha \exp(\beta/T)$. We may substitute the expression for T from (7) into this expression, which gives the bead resistance explicitly as a function of time if the function $W(t)$ is known and independent of the value of the thermistor resistance. This, however, is rarely the case, hence it is necessary to go further and derive another equation in which the dependence of $W(t)$ upon the thermistor resistance may be introduced. To do this we first take the natural logarithm of the resistance-time function, and then solve it for the integral, which gives:

$$\frac{1}{m\sigma}\int_0^t W(t) \exp\left(\frac{Kt}{m\sigma}\right) dt = \frac{\beta \exp\left(\frac{Kt}{m\sigma}\right)}{\ln \frac{R(t)}{\alpha}} - T_0 \exp\left(\frac{Kt}{m\sigma}\right) - c. \qquad (8)$$

This expression may be differentiated with respect to time, and simplified, giving:

$$\frac{1}{m\sigma} W(t) = \frac{K\beta}{m\sigma \ln \frac{R(t)}{\alpha}} - \frac{\beta \frac{dR(t)}{dt}}{R(t) \ln^2 \frac{R(t)}{\alpha}} - \frac{KT_0}{m\sigma}. \qquad (9)$$

In this form the dependence of $W(t)$ may be easily introduced. In fact, if the thermistor is connected to a generator having resistance R_g and an open-circuit voltage $e(t)$ the latter being a known function of time, $W(t)$ may be written

$$W(t) = e^2(t) \frac{R(t)}{[Rg + R(t)]^2}. \qquad (10)$$

Substituting this expression for $W(t)$ into (9) and simplifying gives:

$$\frac{dx}{dt} - \frac{K}{m\sigma} x + \frac{KT_0}{\beta m\sigma} x^2 = -\frac{e^2(t)}{\beta m\sigma R_g}\left[\frac{2x}{\cosh\left(\frac{x}{2} - \frac{\ln \frac{R_g}{\alpha}}{2}\right)}\right]^2 \qquad (11)$$

where $x = \ln R(t)/\alpha$.

This, then, is the complete differential equation for the time behavior of the resistance of a directly heated thermistor. Since it is nonlinear and not one of the standard forms for which an exact solution is known, it will be necessary either to use linear approximations or some method, such as perturbations, in order to obtain the desired result. Exactly which one of the various approximation methods should be used will depend upon the particular problem to which the solution is to be applied.

Conclusion

Equations have been developed which relate the electrical behavior of thermistors to the various thermal and physical constants of the units. These equations define the steady-state performance of the thermistor when connected either in series or parallel in a resistance network. The complete differential equation describing the time behavior of a directly heated thermistor has been developed in a form which may be solved by various linear approximations appropriate to the desired result. These equations have been very useful in the design of control networks employing thermistors.

Acknowledgments

The authors gratefully acknowledge the assistance of many of their colleagues in the Bell Telephone Laboratories. To list all is impossible because of space limitations, but particular mention should be given C. B. Green and G. L. Pearson for preparation of samples, and to H. S. Black, under whose direction these studies made.

Appendix

Given

$$R = a \exp(\beta/T) \qquad (12)$$

then

$$R_0 = a \exp(\beta/T_0)$$

and

$$\ln R/R_0 = \frac{\beta[T_0 - T]}{T_0 T},$$

but we assume $T - T_0 = KP$, hence

$$\ln R_0/R = \frac{-P}{\frac{T_0 P}{\beta} + \frac{T_0^2}{K\beta}} = \frac{-P}{SP + P_0}. \tag{13}$$

In the case of a directly heated thermistor, $R = E/I$ and $P = EI$ where E is the potential drop across the thermistor, and I is the current through it. Substituting this expression in (13) we obtain

$$\frac{E}{I} = R_0 \exp\left[\frac{-EI}{SEI + P_0}\right], \tag{14}$$

and if we make the linear transformation

$$E = v\sqrt{R_0 P_0}$$
$$I = u\sqrt{P_0/R_0}$$

then

$$\frac{v}{u} = \exp\left[\frac{-uv}{Suv + 1}\right]. \tag{15}$$

Starting from (15) we obtain by differentiation

$$\frac{1}{u}\frac{dv}{du} - \frac{v}{u^2} = -\frac{\left[u\frac{dv}{du} + v\right]}{[1 + Suv]^2} \exp\left[\frac{-uv}{1 + Suv}\right],$$

and solving for dv/du gives

$$\frac{dv}{du} = \frac{R_{AC}}{R_0} = \frac{v}{u}\left[\frac{1 - \frac{uv}{[1 + Suv]^2}}{1 + \frac{uv}{[1 + Suv]^2}}\right].$$

Making the substitution $(v/u) = (R_b/R_0) = \epsilon^{-x}$ we obtain.

$$\frac{dv}{du} = \frac{R_{AC}}{R_0} = \epsilon^{-x}\frac{1 - x(1 - Sx)}{1 + x(1 - Sx)}. \tag{16}$$

The resistance of a thermistor is given by

$$\frac{R_b}{R_0} = \exp\left(\frac{-EI}{S(EI) + P_0}\right).$$

if the thermistor is in series with a resistance R, this becomes

$$\frac{R_b}{R_0} = \exp - \left(\frac{I(E - IR)}{SI(E - IR) + P_0}\right).$$

but

$$E - IR = \frac{ER_b}{R + R_b}, \qquad I = \frac{E}{R + R_b}$$

$$\ln \frac{R_b}{R} = -\frac{\frac{E^2 R_b}{(R + R_b)^2}}{P_0 + \frac{SE^2 R_b}{(R + R_b)^2}}.$$

Solving for E^2 and making the substitution $x = \ln R_0/R_b$ we obtain

$$\frac{E^2}{P_0 R_0} = \frac{\left(\frac{R}{R_0} + \epsilon^{-x}\right)^2}{\epsilon^{-x}} \frac{x}{1 - Sx}.$$

Now at the extreme voltages we have $R_{AC} = -R$, substituting the value of $-R_{AC}/R_0$ in terms of x from equation (16) for R/R_0 gives:

$$\frac{E^2}{P_0 R_0} = \frac{\left(-\epsilon^{-x}\frac{1 - x(1 - Sx)}{1 + x(1 - Sx)} + \epsilon^{-x}\right)^2}{\epsilon^{-x}} \frac{x}{1-Sx}$$

$$= \frac{x\epsilon^{-x}}{1 - Sx}\left[\frac{2x(1 - Sx)}{x(1 - Sx) + 1}\right]^2. \tag{17}$$

Differentiating (16)

$$\frac{d\frac{R_{AC}}{R_0}}{dx} = \epsilon^{-x}\left(\frac{-1 + x(1 - Sx)}{1 + x(1 - Sx)} - \frac{-2(1 - 2Sx)}{[1 + x(1 - Sx)]^2}\right).$$

At the maximum negative value this must vanish, and since neither ϵ^{-x} nor $1/[1 + x(1 - Sx)]^2$ vanishes for any finite x we must have:

$$x^2(1 - Sx)^2 - 1 - 2(1 - 2Sx) = 0.$$

Solving for S gives:

$$S = \frac{1}{x} - \frac{2}{x^3} \pm \sqrt{\frac{(x^2 - 2)^2}{x^6} - \frac{x^2 - 3}{x^4}}$$

$$= \frac{x^2 - 2}{x^3}\left(1 \pm \frac{\sqrt{4 - x^2}}{x^2 - 2}\right).$$

Since we know that the actual values of S are small, the negative sign must be used.

$$S = \frac{x^2 - 2}{x^3}\left(1 - \frac{\sqrt{4 - x^2}}{x^2 - 2}\right). \tag{18}$$

Radio Propagation Variations at VHF and UHF*

KENNETH BULLINGTON†, ASSOCIATE, IRE

Summary—**The variations of received signal with location (shadow losses) and with time (fading) greatly affect both the usable service area and the required geographical separation between co-channel stations. An empirical method is given for estimating the magnitude of these variations at vhf and uhf. These data indicate that the required separation between co-channel stations is from 3 to 10 times the average radius of the usable coverage area, and depends on the type of service and on the degree of reliability required. The application of this method is illustrated by examples in the mobile radiotelephone field.**

Introduction

The propagation of radio waves is subject to many variations that are not completely understood. Even over relatively smooth earth the median field intensity measured at distances of 200 miles or more may be 40 to 60 decibels greater than the values predicted by the smooth earth theory. In addition to this "discrepancy" in median values, the field intensity at any given location varies with time, and the resulting fading range tends to increase with both distance and frequency. Although these variations are relatively unimportant at distances within the reliable coverage area, they are of prime importance in estimating the geographical separation that is required between two transmitters assigned to the same frequency.

Other variations in field intensity result from the presence of hills, trees, and buildings, and these are important in estimating reliable coverage areas as well as the interference area. In this case, the measured signals may be 40 decibels or more below the value indicated by the smooth earth theory. Thus it appears that the median field intensity is generally lower at short distances and higher at long distances than is predicted by the smooth earth theory. This situation is unfortunate, since it means that a given frequency assignment cannot be repeated as often geographically as the theory would indicate.

Since the range of the propagation variations may be many tens of decibels, it is obvious that an accurate prediction of the field intensity at a particular location is not possible. The best that can be done is to state the median value to be expected and the probable deviation from the median. It also follows that individual spot measurements may vary over a very wide range, and are of little value in establishing a general trend. In order to determine a general rule for fading, it is desirable to have measurements over a relatively smooth path for many months, preferably a year. Similarly, in order to determine the effects of terrain irregularities, it is necessary to have measurements at a sufficiently large number of locations within 20 or 30 miles from the transmitter where atmospheric fading can usually be neglected.

Some of the general trends in the available data of vhf and uhf that seem suitable for statistical analysis are summarized in the following sections. The results are applied to the problem of estimating the required separation between two co-channel transmitters in the 40- to 50-Mc range. This paper does not consider sky-wave propagation, although ionospheric reflections can occur at frequencies above 30 Mc.

Differences Between Measured Field Intensity and Values Predicted by Smooth Earth Theory

The theory of radio propagation over plane earth is well established. It reduces to the familiar concept of a direct and reflected ray for antenna heights greater than about one wavelength above ground. The smooth earth theory which provides the correction for the curvature of the earth has not been checked as completely as the plane earth theory.[1–3] It agrees reasonably well with experimental results as long as the correction for earth's curvature is less than 20 or 30 decibels but it seems to fail when the indicated curvature loss is greater than 30 to 40 decibels. This means that theoretical curves based on the smooth earth theory may be in serious error at distances beyond 100 miles for 30 to 40 Mc, beyond 60 miles for 300 Mc, and beyond 30 miles for 3,000 Mc, unless the antenna heights are sufficiently high to approximate plane earth conditions.

Most of the data on long range ground wave transmission are at frequencies in the 40- to 50-Mc range.[4–6] The field intensities for the ten paths listed in the following table are the median values that were recorded for periods of at least six months to a year. The measured values have been adjusted to an effective radiated power of 1 kw.

The right-hand column in Table I shows the approximate median field intensities that would be expected if the antenna heights were 500 and 30 feet instead

* Decimal classification: R113.2. Original manuscript received by the Institute, April 4, 1949. Presented, 1949 IRE National Convention, New York, N. Y., March 8, 1949.

† Bell Telephone Laboratories, Inc., New York, N. Y.

[1] C. R. Burrows and M. C. Gray, "The effect of the earth's curvature on ground-wave propagation," Proc. I.R.E., vol. 29, pp. 16–24; January, 1941.

[2] K. A. Norton, "The calculation of ground-wave field intensity over a finitely conducting spherical earth," Proc. I.R.E., vol. 29, pp. 623–639; December, 1941.

[3] K. Bullington, "Radio propagation at frequencies above 30 Mc," Proc. I.R.E., vol. 35, pp. 1123–1136; October, 1947.

[4] TID Report 2.4.5., "Summary of tropospheric propagation measurements and the development of empirical propagation charts," Fed. Communications Commission (27989); October 20, 1948.

[5] G. S. Wickizer and A. M. Braaten, "Propagation studies at 45.1, 474, and 2,800 Mc," Proc. I.R.E., vol. 35, pp. 670–680; July, 1947.

[6] G. N. Packard and H. T. Stetson, "Tropospheric reception at 42.8 Mc and meteorological conditions," Proc. I.R.E., vol. 35, pp. 1445–1450; December, 1947.

TABLE I

Path No. Used in T.I.D. 2.4.5.	Miles	MC	Measured Median Field db above $1\mu V/m$	Duration of Measurement Months	Antenna Heights		Median Field Adjusted for Ant. Hts. of 500′ & 30′ db above $1\mu V/m$
					Trans. Feet	Rec. Feet	
	42.5	45.1	49.0	12	1,300	80	32
34	45.0	47.1	30.6	12	570	50	25.5
35	68.0	47.1	18.0	6	570	30	17.0
	70.0	45.1	31.0	12	1,300	132	10.0
3	104.0	45.7	7.6	7	380	30	10.0
4	122.0	45.5	3.0	10	690	50	− 4.0
31	167.0	42.8	4.1	11	700–900	50	− 4.0
36	186.0	47.1	− 6.2	12	570	30	− 7.2
1	198.0	42.8	−11.7	11	700–900	30	−15.7
2	337.0	44.3	−31.0	11	1,600–680	30	−41 −33.5

of at the heights shown. The corrections are based on the plane earth theory which indicates that the field intensity is increased 6 decibels when the antenna height is doubled.[7] The corrected values are plotted on Fig. 1.

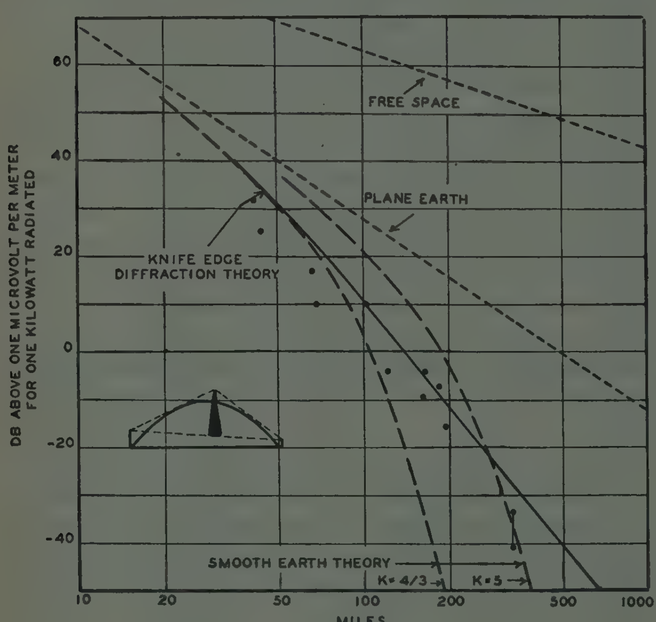


Fig. 1—Measured and theoretical field intensities in the 40- to 50-Mc range, adjusted for a radiated power of 1 kw and antenna heights of 500 and 30 feet.

The theoretical curve based on the smooth earth theory, including the average refraction in the earth's atmosphere, is indicated by the dashed line for $k=4/3$, where k is the ratio of the effective earth's radius to the true earth's radius. It will be noted that the experimental values of field intensity decrease much more gradually with increasing distance than is indicated by the smooth earth theory. Any assumed change in the value of k moves the theoretical curve horizontally (and slightly downward) without changing its general shape. This is indicated by the dashed line for $k=5$. The assumption of another value of k would not provide a better fit with the experimental data, and in addition a long-term average value of k that is appreciably greater than 4/3 would be difficult to justify from a meteorological standpoint. In a similar manner, a change in the conductivity of the ground over the extreme range from perfectly insulating to perfectly conducting affects the theoretical values in approximately the same manner as a change from $k=4/3$ to $k=2$. There is no apparent method of adjusting the smooth earth theory by changes in the values of k or of the ground constants to fit the available experimental data at 45 Mc.

The solid line on Fig. 1 is a theoretical curve obtained by assuming that the field intensity is below the plane earth value by the diffraction loss over a knife edge whose height is determined by the intersection of two straight lines drawn tangent to the earth's surface from the two antennas. It assumes the average refraction represented by $k=4/3$. The best empirical curve drawn through these experimental points has the same general shape as the solid theoretical curve, but is lower by about 6 dbs. Thus the knife-edge theory, which should not be applicable, seems to give a better correlation with the experimental data than the smooth spherical earth theory. In addition, the apparent failure of the smooth earth theory is most noticeable in the region where it should be most accurate, since the fundamental assumption of a smooth sphere is more closely approached as the distance is increased.

The long-term experimental data at higher frequencies are too meager for firm conclusions, but the present results indicate that the median field intensity at points beyond the line of sight is also higher than predicted by the smooth earth theory, but is less than indicated by the knife-edge diffraction theory. Whether the smooth earth theory holds for microwaves must await long-term recordings at several points far beyond the optical horizon.

[7] According to the smooth earth theory, this method is in error at 45 Mc for points beyond the horizon by about 0.7 db at 500 feet, 3 db at 1,000 feet and 5 db at 1,500 feet, but these differences are small compared with the discrepancy between the experimental and theoretical results.

Effects of Atmospheric Fading

The variations in field intensity around the median value are presumably the result of changes in the atmospheric conditions. The field intensity tends to be higher in summer than in winter and higher at night than during the day for over-land paths that are beyond the optical line of sight. The fading range ordinarily increases with an increase in either distance or frequency. As a first approximation, the distribution of field intensities follows a normal probability law in decibels, except that the signal seldom exceeds the value expected over plane earth.

The difference in decibels between the values exceeded 10 per cent of the time and the median values are shown on Fig. 2 for the 40- to 50-Mc paths for which

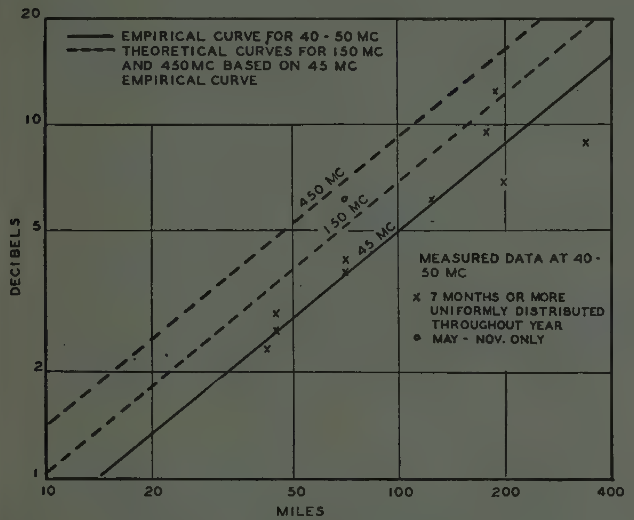


Fig. 2—Atmospheric fading; ratio of field intensity exceeded 10 per cent of time to field intensity exceeded 50 per cent of time.

data are available for more than six months. It will be noted that considerable spread exists between the measured values and any single line drawn to represent the data. Most of this spread probably results from factors other than distance such as antenna height, type of terrain, and different climatic conditions.

At frequencies higher than 40- to 50-Mc, the long-term data on atmospheric fading are meager and an empirical relationship drawn from a small number of points may be misleading. However, the available data at frequencies below 500 Mc are not inconsistent with the dashed lines on Fig. 2 for 150 and 450 Mc. These curves are derived from the 40- to 50-Mc data by assuming that the fading range varies with frequency in the same manner as the diffraction loss introduced by the earth's curvature. The diffraction theory (for either a smooth sphere or the knife edge shown on Fig. 1) indicates that the diffraction loss varies as the product of the frequency times the cube of the distance. This extension of the 40- to 50-Mc data by theoretical considerations cannot be carried too far. For example, if this theory were assumed to hold at 4,000 Mc, the resulting estimated fading range would be greater than the measured values on good optical paths, and would be less than the measured values on paths where the receiver is far below the optical horizon.

Some atmospheric fading also occurs over good optical paths.[6,8] For one per cent of the time at 45 Mc the field intensity is about 2 or 3 db less than the median value, while at 4,000 Mc it may be 15 to 20 db below the median value.

Effect of Hills

The available data on the effect of hills indicate that the shadow losses increase with frequency and with the roughness of the terrain.[9] The variations in shadow losses around the median value usually follow a normal probability law in decibels.

The principal experimental data on the effect of hills are shown on Fig. 3. The roughness of the terrain is

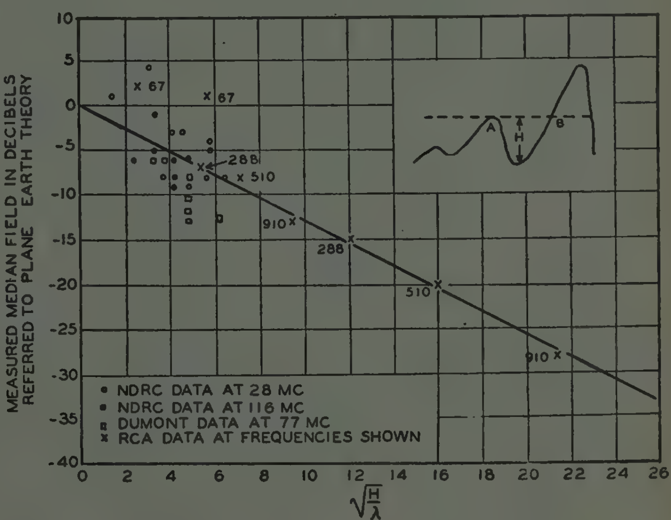


Fig. 3—Effects of terrain on radio propagation; radio of measured median field intensities to the plane earth theoretical values for random locations between points A and B.

assumed to be represented by height H shown on the profile at the top of the drawing. This height is the difference in elevation between the bottom of the valley and the elevation necessary to obtain line of sight with the transmitting antenna. The difference between the measured values of field intensity and the values to be expected over plane earth is computed for each point of measurement between A and B. The median value for each general location is plotted on Fig. 3 as a function of $\sqrt{H/\lambda}$. The choice of this parameter is based on knife-edge diffraction theory which indicates that the

[8] A. L. Durkee, "Results of microwave propagation tests on a 40-mile overland path," Proc. I.R.E., vol. 36, pp. 197–205; February, 1948.

[9] G. H. Brown, J. Epstein, and D. W. Peterson, "Comparative propagation measurements; Television transmitters at 67.25, 288, 510 and 910 megacycles," *RCA Rev.*, vol. 9, pp. 177–201; June, 1948.

parameter should be $\sqrt{H/\lambda}\tan\theta$, where θ is the angle of bend around the obstruction, and it is assumed that the variation in θ is much less than the variation in H/λ.

The open circles indicate data taken at 28 Mc and the solid circles are data at 116 Mc. The crosses indicate data at 67, 288, 510 and 910 Mc. It will be noted that there is considerable spread, especially in the region where the shadow loss is less than 10 db. The solid line is an empirical approximation for the median shadow losses. In a similar manner an estimate of the shadow losses exceeded in 10 per cent of the possible locations can be obtained.

These empirical relationships are summarized in the nomogram shown in Fig. 4. The scales on the right-hand line indicate both the median value of shadow loss

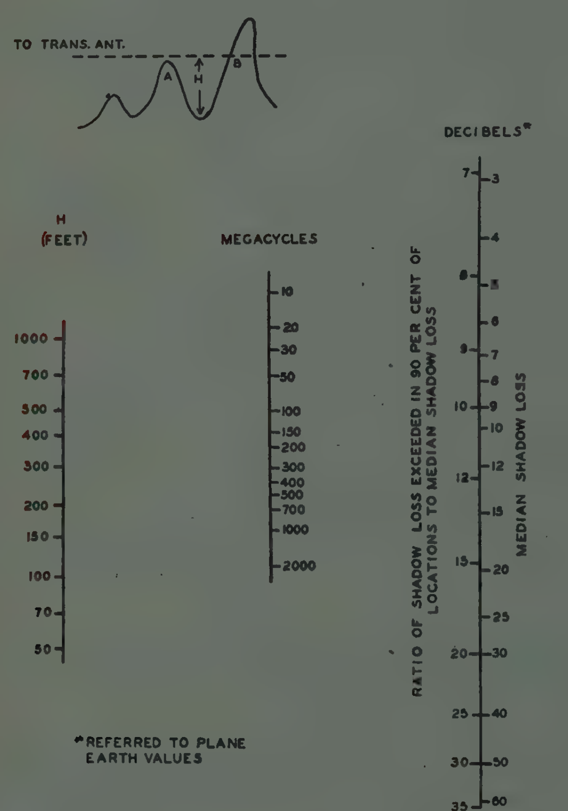


Fig. 4—Estimated distribution of shadow losses for random locations between points A and B.

and the difference in shadow loss to be expected between the median and the 10 per cent values. For example, with variations in terrain of 500 feet, the estimated median shadow loss at 450 Mc is about 20 db and the shadow loss exceeded in only 10 per cent of the possible locations is about $20+15=35$ db. It will be recognized that this analysis is based on large-scale variations in field intensity, and does not include the standing wave effects which sometimes cause the field intensity to vary considerably in the matter of a few feet.

Effects of Buildings and Trees

The shadow losses resulting from buildings and trees follow different laws than those caused by hills. Buildings are more transparent to radio waves than the solid earth, and there is ordinarily much more back scatter in the city than in the open country. Both of these factors tend to reduce the shadow losses caused by the buildings but, on the other hand, the angles of diffraction over or around the buildings are usually greater than for natural terrain. In other words, the artificial canyons caused by buildings are considerably narrower than natural valleys, and this factor tends to increase the loss resulting from the presence of buildings. The quantitative data on the effects of buildings are confined almost entirely to New York City. These data seem to indicate that in the range of 40 to 450 Mc there is no significant change with frequency, or at least the variation with frequency is somewhat less than the square root relationship noted in the case of hills. The median field intensity at street level for random locations in Manhattan (New York City) is about 25 db below the corresponding plane earth value. The corresponding values for the 10 per cent and 90 per cent points are about 15 and 35 db, respectively.

Trees and other objects whose dimensions are of the same order of magnitude as the wave length (at vhf) tend to act as parasitic radiators. The average shadow loss is small at 30 to 50 Mc, but increases with increasing frequency. The distribution of losses in the immediate vicinity of trees does not follow a normal probability law but is more accurately represented by Rayleigh's law, which is the distribution of the sum of a large number of equal vectors having random phases. In this type of distribution the 10 per cent and 1 per cent points are about 8 and 18 decibels, respectively, below the median value.

Application to Co-channel Interference Problem

In order to make the most efficient use of the radio-frequency spectrum, it is necessary to repeat frequency assignments as closely as possible without producing unreasonable co-channel interference. The amount of cochannel interference that can be tolerated depends on the type of receiver, and on the quality and reliability of the service required. For example, with FM receivers having a 5:1 deviation ratio, a signal to interference ratio of 4 to 10 db or more is required depending on the quality desired, while for television broadcasting the FCC standards specify a signal to interference ratio of 40 db.

If the variations in radio transmission could be ignored, it would be possible to determine an optimum geographical layout from the required signal to interference ratio and a curve of median field intensity versus distance, such as Fig. 1. However, the magnitude of the transmission variations may be comparable to or

greater than the signal to interference ratio required by the receiver and hence these variations cannot be ignored. It is still possible to use the transmission curve of median field intensities versus distance, providing that the required signal to interference ratio is taken as the above value plus an allowance for the magnitude of variations to be expected and the reliability desired. The principal variations seem to follow the normal probability law, and the sum of four independent normal distributions indicates that the required signal-to-interference ratio S'/I' is given by

$$20 \log \frac{S'}{I'} = 20 \log \frac{S}{I} + K\sqrt{X_S{}^2 + Y_F{}^2 + Y_S{}^2 + X_F{}^2},$$

where

S/I = signal-to-interference ratio required by the receiver for the quality of service desired

X_S = difference in decibels between the 10 per cent and 50 per cent values of shadow losses for the desired signal

X_F = same for atmospheric fading

Y_F = difference in decibels between the 10 per cent and 50 per cent values of atmospheric fading for the undesired signal

Y_S = same for shadow losses

K = 0 for 50 per cent reliability (based on normal probability law)

= 1 for 90 per cent reliability

= 1.8 for 99 per cent reliability.

Since the variations are not entirely independent, some judgment is needed in the use of the above formula. The two principal variations are the shadow loss distribution for the desired signal X_S (shown on Fig. 4) and the atmospheric fading distribution for the undesired signal Y_F (shown on Fig. 2). The shadow loss distribution for the undesired signal X_F cannot be taken at full value, since the local terrain at the receiver may affect both signals alike; but, also, it cannot be neglected, since the two signals are usually coming from nearly opposite directions. The most probable value of Y_S is assumed in the following example to be one-half of the value shown in Fig. 4. In a similar manner the atmospheric fading of the desired signal is not independent of the fading of the undesired signal, that is, there is some tendency for the two signals to fade up and down together. Fortunately in this case, the value of X_F is always small compared with Y_F (and also usually small compared with X_S), so that it can be neglected without an appreciable effect on the final results.

As an example, consider a mobile radiotelephone system operating at 40 Mc over terrain having variations of 200 feet. It is assumed that the interfering transmitter has the same effective radiated power and antenna height as the desired transmitter, and that a signal-to-interference ratio of at least 6 db is desired for more than 90 per cent of the locations (and time) at the edge of the coverage area.[10] The value of X_S is shown on Fig. 4 to be about 7.5 db and the value of Y_F is shown on Fig. 2 to be about 3 db at 50 miles, 5 db at 100 miles, and 9 db at 200 miles. It follows that the value of 20 log S'/I' for a distance of 100 miles from the receiver to the undesired transmitter is given by

$$20 \log \frac{S'}{I'} = 6 + \sqrt{(7.5)^2 + (5)^2 + \left(\frac{7.5}{2}\right)^2 + 0}$$
$$= 15.8 \text{ db}.$$

This means that when the distance from the receiver to the undesired transmitter is 100 miles the reliable coverage radius of the desired transmitter is limited to that distance at which the median field intensity from the desired transmitter is 15.8 db higher than the median field intensity at a distance of 100 miles. The required separation between the two transmitters is then 100 miles plus the coverage radius.

The principal remaining choice is the proper relationship for the median field intensity versus distance. Although the absolute values of field intensity depend on the antenna heights and radiated power, these factors are equal for both the desired and undesired signals, so only the slope of the curve is important in this example. For distances between 20 and 250 miles it is assumed

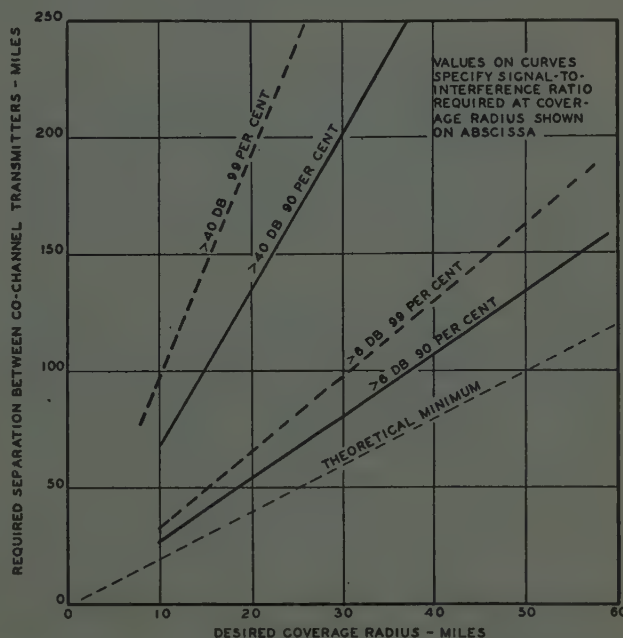


Fig. 5—Estimated separation required between co-channel transmitters in 40- to 50-Mc range, assuming 200-foot terrain irregularities.

[10] This example is based on two frequency operations, and considers the interference into a mobile receiver from an unwanted fixed transmitter. For single frequency operation, the principal co-channel interference occurs from a fixed transmitter into a fixed receiver, and the required separation between co-channel systems is considerably greater than shown in this example.

that the signal intensities are given by the solid line on Fig. 1. In the above example the median signal intensity at 60 miles is about 16 db stronger than at 100 miles. Thus the coverage radius is 60 miles, and the required separation between co-channel stations is $100+60=160$ miles.

In a similar manner, additional points can be obtained for other distances between the receiver and the unwanted transmitter. The resulting separation required between co-channel transmitters is shown on Fig. 5 by the solid line marked ">6 db, 90%." Thus for a coverage radius of 20 miles the separation required at 45 Mc is about 55 miles. The corresponding value for 99 per cent reliability is about 67 miles. Similar values for a required signal to interference ratio of 40 db at the receiver are also shown.

When higher frequencies are used or when more rugged terrain is considered, the required separation between co-channel stations will be greater than shown on Fig. 5. When the radiated powers or antenna heights are not the same for the two transmitters, the first-order effect is for the stronger station to gain coverage area at the expense of the weaker station (assuming equal receiving antenna heights) without any substantial change in the required separation between stations.

It will be realized that the coverage radii shown on Fig. 5 are the maximum values to be expected for a given separation, since they are based on co-channel interference only. The actual coverage radius may be less than shown if the signal level is insufficient to over-ride the noise at this distance. For optimum results it is obvious that a proper balance is required between the limitations set by noise and those imposed by co-channel interference.

The Dynamic Sensitivity and Calibration of Cathode-Ray Oscilloscopes at Very-High Frequencies*

HANS E. HOLLMANN†, ASSOCIATE, IRE

Summary—**The drop in sensitivity of a cathode-ray oscilloscope at very-high frequencies is investigated through the extension of the simple inversion formula by introducing the exit displacement as well as the additional deflection in the stray fields. Since the actual stray field distribution cannot be predicted by any analysis, it is determined on the basis of a static or quasi-static calibration. The practical use of the somewhat complex formulas is illustrated by several examples.**

THE GENERAL BEHAVIOR of cathode-ray tubes at very-high frequencies, where the deflection is influenced by transit-time effects, appears to have received little attention up to now. The only worker who used the well-known basic formula for dynamic sensitivity[1-3] in evaluating the reduction in sensitivity of his three-beam micro-oscillograph at increasing frequency is Lee.[4] In view of the fact that this instrument facilitates experimental investigation in the microwave range, in addition to transit-time oscillography,[5] it may be of interest to consider the dynamic deflection of an electron beam in greater detail.[6]

The relative dynamic sensitivity P_ω of a cathode-ray tube, whose beam is deflected between two parallel plates, is the ratio of the dynamic sensitivity A_ω at any very-high frequency to the corresponding static or quasi-static sensitivity A_0 at direct or relatively low-frequency alternating voltages:

$$P_\omega = \frac{A_\omega}{A_0} = \frac{\sin\frac{\Phi}{2}}{\frac{\Phi}{2}}, \tag{1}$$

where Φ is the transit-time angle required for the electrons to pass the deflecting field. By introducing the electron velocity v_0 corresponding to the kv plate potential V_p, and by expressing the frequency by the wavelength λ, the transit-time angle, referred to the

* Decimal classification R201.7×R371.5. Original manuscript received by the Institute, July 6, 1948; revised manuscript received, September 6, 1949.

† Oxnard, Calif.

[1] R. Mines, "Measurement in electrical engineering by means of cathode rays," *Jour. IEE* (London), vol. 63, pp. 1096–1098; 1925.

[2] Hans E. Hollmann, "Die Braunsche Röhre bei sehr hohen Frequenzen" (The cathode-ray tube at very-high frequencies) *Zeit. fuer Hochfrequenz.*, vol. 40, pp. 97–103; 1932.

[3] Hans E. Hollmann, "The use of the cathode-ray oscilloscope at ultra-high frequencies," *Wireless Eng.*, vol. 10, pp. 430–433 and 484–486; 1933.

[4] Gordon M. Lee, "A three-beam oscillograph at frequencies up to 10,000 megacycles," Proc. I.R.E., vol. 34, pp. 121–127; March, 1936.

[5] Hans E. Hollmann, "Ultra-high-frequency oscillography," Proc. I.R.E., vol. 28, pp. 213–219; May, 1940.

[6] Hans E. Hollmann, "Das Verhalten der Kathodenstrahlröhre im Laufzeitgebiet" (The behavior of the cathode-ray tube in the transit-time range), *Fort. der Hochfrequenz.*, vol. 1, pp. 453–486; 1941.

actual length l_p of the plates, becomes

$$\Phi_p = \frac{\omega l_p}{v_0} = 31.6\pi \frac{l_p}{\lambda\sqrt{V_p}}. \tag{2}$$

The solid curve in Fig. 1 shows the corresponding inversion relation between P and Φ_p. At multiples of 2π, the

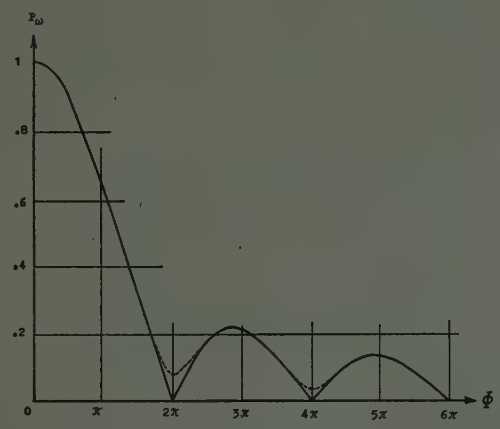


Fig. 1—The relative dynamic sensitivity P_ω in relation to the transit-time angle Φ.

oscilloscope becomes ineffective, but between these zero points subsequent maxima with decreasing amplitudes appear until the oscilloscope remains completely ineffective at $\Phi_p \to \infty$.

At small values of Φ_p (1) may be developed into a series and yields a percentage error of

$$p_{0/0} = -\frac{4\cdot 10^4}{V_p}\left(\frac{l_p}{\lambda}\right)^2. \tag{3}$$

For example, at $V_p = 50$ kv, $l_p = 0.2$ inch $= 0.5$ cm, and $\lambda = 10$ cm, the error is about -2.1 per cent.

The basic formula (1) neglects two facts: first, the displacement of the electrons leaving the deflecting field, and second, the additional deflection in the stray fields. For a more accurate analysis, the basic formula (1) must be extended.

Usually, the deflection of a cathode ray is based only on the deflecting angle α, as shown in the upper part of Fig. 2, and the deviation of the luminous spot at the fluorescent screen is then given by $D = L \tan \alpha$ where L denotes the length in centimeters of the free beam lever equal to the distance between plates and screen.

Besides this angular deflection, there exists an exit displacement y_e so that the total deviation of the spot becomes $D = L \tan \alpha + y_e$. As a result, the static or quasi-static deflecting sensitivity under the influence of a direct- or low-frequency alternating peak voltage V_0 in volts across the plates with the separation a becomes

$$A_0 = \frac{D}{V_0} = \frac{l_p}{2aV_0}\left(L + \frac{l_p}{2}\right). \tag{4a}$$

The equation discloses the well-known fact that the parabolic track of the beam inside the deflecting field may be replaced by the dotted backward extension of the external beam lever in Fig. 2, cutting the axis at the center of the field. Hence, the static or quasi-static deflection may be visualized as caused by a linear beam of the length $L + l_p/2$ deviating from the axis at the center of the field by the angle α.

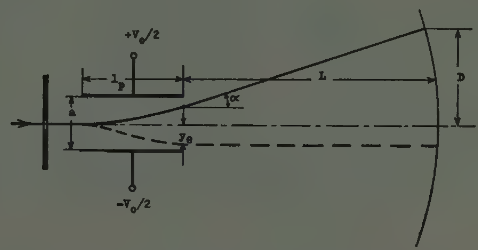


Fig. 2—Schematic representation of a statically and dynamically deflected electron beam.

A similar analysis yields the dynamic sensitivity

$$A_\omega = \frac{e}{m}\frac{1}{a\omega^2}\left\{2(1-\cos\Phi) + \Phi^2 - 2\Phi\sin\Phi + 2\Phi^2\frac{L}{l}\left(1+\frac{L}{l}\right)(1-\cos\Phi)\right\}^{1/2}. \tag{4b}$$

At transit-time angles $n2\pi$ ($n = 1, 2, 3 \cdots$), the angular deflection disappears and (4b) yields only the dynamic exit displacement y_e. In this peculiar case, the beam is displaced only parallel, as may be seen by the dashed track in the lower part of Fig. 2. Dividing (4b) by (4a) yields the extended inversion formula

$$P_\omega = \frac{\left\{2(1-\cos\Phi) + \Phi^2 - 2\Phi\sin\Phi + 2\Phi^2\frac{L}{l}\left(1+\frac{L}{l}\right)(1-\cos\Phi)\right\}^{1/2}}{\Phi^2\left(\frac{L}{l}+\frac{1}{2}\right)}. \tag{4c}$$

At transit-time angles $\Phi = n2\pi$, (4c) is simplified into the relative exit displacement

$$P_0 = \frac{1}{\Phi\left(\frac{L}{l}+\frac{1}{2}\right)}. \tag{4d}$$

The dotted curve in Fig. 1 shows the deviation of the extended (4c) from the simple (1) for a quotient $L/l_p=2$. Obviously, the extension becomes noticeable only in the vicinity of the zero points leveling them into minima, and the more L/l_p decreases, the flatter they become.

For the sake of completeness, it may be mentioned that the present formulas give the peak deflection because no entrance phases can be observed. Only at supercritical voltages V_0 the beam temporarily strikes the plates, and certain entrance phases are cut off. As a result, the deflecting line on the screen is interrupted as was shown by the overmodulation spectrum.[5]

In order to introduce the second effect, i.e., the additional deflection in the stray fields, it is convenient to approximate the actual stray-field distribution, the solid curves in Fig. 3, by the dashed cubic parabolas. The

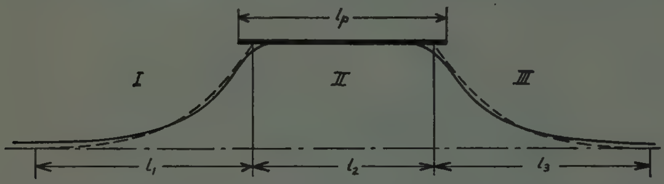


Fig. 3—The stray fields of an electrostatic deflection field. The solid curve shows the actual stray fields, whereas the dotted curves represent a first-order approximation by cubical parabolas.

electrons then pass three different fields whose field-strength distribution may be expressed

Entrance stray field I: $\dfrac{V_0}{a}\left(\dfrac{x}{l_1}\right)^3$;

Homogeneous center field II: $\dfrac{V_0}{a}$;

Exit stray field III: $\dfrac{V_0}{a}\left(\dfrac{x-(l_1+l_2+l_3)}{l_3}\right)^3$;

with their respective lengths

$$l_1 = l_3 = 3.1a \quad \text{and} \quad l_2 = l_p - a/2. \tag{5}$$

Converting all three fields into one rectangular field having the same area yields an effective field having the length

$$l_{\text{eff}} = l_p + 1.05a \sim l_p + a, \tag{6a}$$

or the effective format

$$F_{\text{eff}} = F_p + 1.05 \sim F_p + 1. \tag{6b}$$

A comparison of the present approximation with more accurate calculations[7] reveals an error of only a few per cent.

In the transit-time region the cubic distribution must be taken into account, because the stray fields depend on the momentary time as well as on the momentary localization of any electron under consideration. Adding the complex partial deflections in the three individual spaces gives the new equation:

$$P_\omega = \frac{6}{\phi_1^3\left(\dfrac{\phi_1}{2}+\phi_2\right)}\left\{[\phi_1^2 - 2(1-\cos\phi_1)]\cos\frac{\phi_2}{2} + 2[\phi_1 - \sin\phi_1]\sin\frac{\phi_1}{2}\right\} \tag{7a}$$

where

$$\phi_1 = \phi_3 = 100\pi\,\frac{a}{\lambda\sqrt{V_p}},$$

and

$$\phi_2 = 31.6\pi\,\frac{l_{p\text{cm}} - 1/2}{\lambda_{\text{cm}}\sqrt{V_p}}$$

denote the individual transit-time angles. For the sake of simplicity, it may be more convenient to introduce the ratio

$$q = \frac{\phi_1}{\phi_2} = \frac{\phi_3}{\phi_2} = \frac{3.1}{F_p - 1/2} \tag{8}$$

between the transit-time angles in both stray fields on

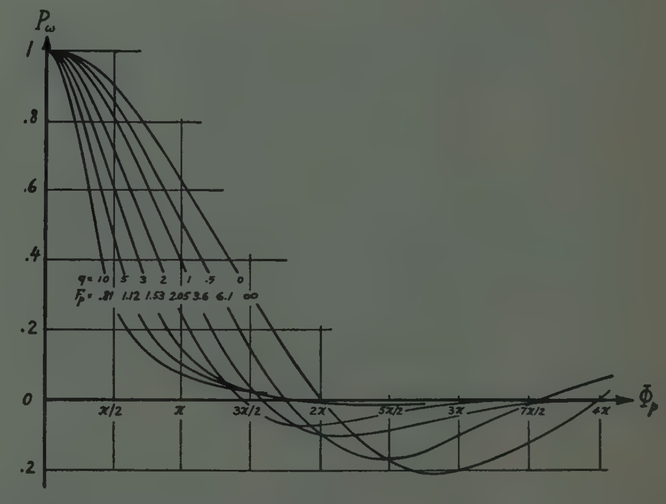


Fig. 4—The relative dynamic sensitivity P_ω in relation to the transit-time angle Φ_p with various parameters q or F_p, respectively.

the one side and in the center field on the other side, so that (7a) takes the form

$$P_\omega = \frac{12}{(q\phi_2)^4\left(\dfrac{2}{q}+1\right)}\left\{[(q\phi_2)^2 - 2(1-\cos q\phi_2)]\cos\frac{\phi_2}{2} + 2[q\phi_2 - \sin q\phi_2]\sin\frac{\phi_2}{2}\right\}. \tag{7b}$$

[7] J. T. McGregor-Morris and V. A. Hughes, "Experimental verification of the theory of the cathode-ray oscillograph," *Jour. IEE* (London), vol. 79, pp. 454–462; 1936.

Fig. 4 shows the relative dynamic sensitivity in relation to the original transit-time angle

$$\Phi_p = \frac{\phi_2}{1 - \frac{1}{2F_p}}$$

with various parameters F_p or q. At $q \rightarrow 0$ or at great values of F_p, respectively, the stray fields vanish and the corresponding extreme curve is identical with the solid curve in Fig. 1. The smaller F_p becomes, the more the stray fields will increase and the relative dynamic sensitivity, therefore, will not only decrease more rapidly at the beginning, but the subsequent inversion maxima become flatter indicating that the inversion effects become more and more blurred. In the extreme $q \rightarrow \infty$ the homogeneous center space vanishes completely leaving only both stray fields. For example, this occurs if the deflecting field expands between two balls or wires. Then (7b) for great values of ϕ_1 is simplified into

$$P_\omega \rightarrow \frac{12}{\phi_1^2},$$

that is, the relative dynamic sensitivity finally approaches zero.

In order to achieve, under all circumstances, the highest accuracy, the previous calculations must be combined with the exit displacement effect. Assuming some simplifications, the relative dynamic sensitivity becomes generally

$$\begin{aligned} P_\omega = \frac{1}{\phi_2^2\left[1 + \frac{L}{l_2}\left(\frac{q}{2}+1\right)\right]} \Bigg\{ & \Big[2(1 - \cos\phi_2) \\ & - 2\phi_2 \sin\phi_2 + \phi_2^2\Big] \\ & - \frac{6L}{l_2 q^3 \phi_2^2}\left[2 \sin\frac{\phi_2}{2}\left(N \cos\left(q + \frac{1}{2}\right)\phi_2\right.\right. \\ & \left. - M \sin\left(q + \frac{1}{2}\right)\phi_2\right) \\ & \left. + (q+1)\phi_2\left(M \sin\frac{\phi_2}{2} - N \cos\frac{\phi_2}{2}\right)\right] \\ & + 9\left(\frac{L}{l_2}\right)^2 \frac{M^2 + N^2}{\phi_2^4 q^6}\Bigg\}^{1/2} \end{aligned} \tag{9}$$

with the functions

$$\begin{aligned} M = 2 &+ (q\phi_2)^2[\cos q\phi_2 + \cos(q+1)\phi_2] \\ &+ 2q\phi_2[\sin(q+1)\phi_2 - \sin q\phi_2] \\ &+ 2\cos q\phi_2[\cos(q+1)\phi_2 - 1] \\ &- 2\cos(q+1)\phi_2 - 2\sin q\phi_2 \sin(q+1)\phi_2 \end{aligned}$$

and

$$\begin{aligned} N = (q\phi_2)^2&[\sin q\phi_2 + \sin(q+1)\phi_2] \\ &+ 2q\phi_2[\cos q\phi_2 - \cos(q+1)\phi_2] \\ &+ 2\sin(q+1)\phi_2[\cos q\phi_2 - 1] \\ &+ 2\sin q\phi_2[\cos(q+1)\phi_2 - 1]. \end{aligned}$$

Although these equations are hard to handle, they have been worked out in order to show the accuracy of the analysis in the following abnormal example.

In an actual cathode-ray tube, the stray fields do not expand unrestrictedly and, therefore, their distribution differs from the analytical assumptions. In general, they are limited and influenced by some ground apertures, internal leads, and even the charged glass walls. Since all these effects cannot be covered by the theory, the following procedure may be utilized for the purpose of approximating the actual stray fields as far as possible. A direct- or a low-frequency calibration first gives the actual F_{eff} of the equivalent rectangular field which must be converted into the parabolic representation corresponding to Fig. 3. With the aid of the geometrical F_p and the experimentally obtained F_{eff}, the parameter q can be computed by the formula

$$q = 2\left(\frac{F_{eff}}{F_p - \frac{1}{2}} - 1\right). \tag{10}$$

A specially constructed test tube with $l_p = 3.9$ inch $= 10$ cm, $a = 0.39$ inch $= 1$ cm, $F_p = 10$, and $L = 7.5$ inch $= 19$ cm has a theoretical sensitivity of 1.2 mm/v. The actual sensitivity has been measured as 1.37 mm/v so that

$$F_{eff} = \frac{L}{a}\left[\sqrt{1 + \frac{4A_0'V_p}{L^2}} - 1\right] \tag{11}$$

yields $F_{eff} = 11.1$ and $q = 0.33$. With all these values and the quotient $L/l_2 = 19/(10 - 0.5) = 2$, (9) yields the curve

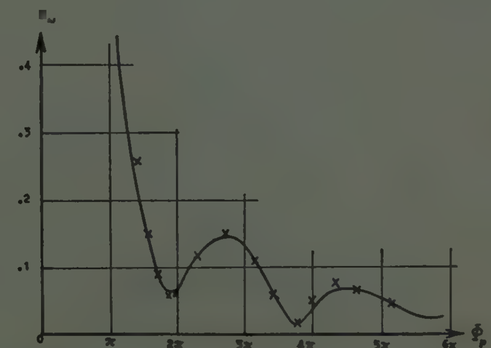


Fig. 5—The relative dynamic sensitivity of a test tube with a field format F_p of 10. The crosses represent points of measurement.

shown in Fig. 5 indicating the relative dynamic sensitivity in relation to the geometrical transit-time angle Φ_p. The crosses are measuring points obtained at 238

Mc at various plate voltages V_p. The accurate high-frequency voltage across the plates has been measured by means of two diodes constructively combined with the deflecting plates in such a manner that the latter form the anodes whereas two little hairpin cathodes approached them at a very small distance.

In respect to the large field format, the extension of the initial (1) concerns more the magnitudes of P_ω than the position of its minima. This picture changes at smaller field formats or at larger stray fields, respectively. A second test tube with $F_p=1.18$ and $F_{eff}=1.8$ has been investigated, indicating that the extended stray fields must be limited in the aforementioned manner. With the parameter $q=3.1$, (7b) produces the solid curve in Fig. 6 which agrees very well with the measuring points obtained at 366 Mc. The dotted and dashed curves reveal the improvement of the extended analysis, namely, the dotted curve shows the course of P_ω corresponding to (1) with Φ_p, and the dashed curve by introducing l_{eff}. Obviously, the deviation becomes more pronounced the more P_ω decreases.

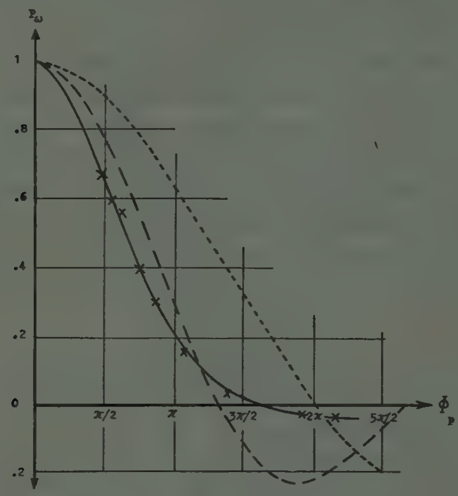


Fig. 6—Calibration curves similar to Fig. 5, but for a test tube with $F_p=1.175$. The solid curve is computed by means of (7b) and the dotted and dashed curves by means of the basic (1), the former utilizing the actual plate length l_p, and the latter with the aid of an effective field format as determined by a static calibration.

As a third example, the dynamic sensitivity of the micro-oscillograph has been computed. Its plates, with $l_p=0.2$ inch and $a=0.08$ inch, have an actual format $F_p=2.5$. The static sensitivity A_0' is given by the manufacturer, at $V_p=50$ kv and with $L=5$ cm, as 2.10^{-4} cm/v, so that (11) gives $F_{eff}=3$, a value which is somewhat larger than the ideal stray field correction for free space. Furthermore, (10) gives $q=1$ and (7b) produces the solid curve in Fig. 7. As a result, the sensitivity de-

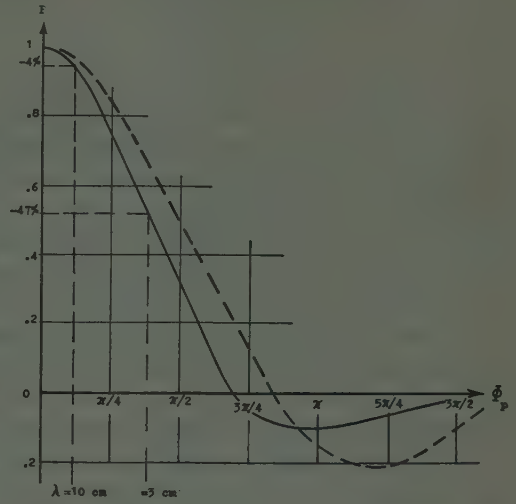


Fig. 7—The dynamic sensitivity of the three-beam micro-oscillograph. The solid curve is computed by the accurate (7b), the dashed curve by (1).

creases, e.g., at 3,000 Mc ($\lambda=10$ cm) by 4 per cent, and at 10,000 Mc ($\lambda=3$ cm) by 47 per cent. If the dynamic sensitivity is computed from the basic equation (1) introducing an effective field length

$$l_{eff} = l_p \frac{F_{eff}}{F_p} = 0.096 \text{ inch} = 0.244 \text{ cm},$$

the dashed curve is obtained giving only a very rough approximation.

Finally, it may be mentioned that the analysis applies only when the stray fields do not change at very-high frequencies with respect to static conditions, e.g., as occurs if the deflecting plates are mounted in a shielding chamber with cavity resonances. Since such unfavorable conditions must be avoided, in view of the vhf impedance coupled to the plates, the practical significance of the present analysis and calibration method will not be lessened.

Acknowledgments

The author wishes to extend his sincere thanks to Gordon M. Lee of the Central Research Laboratories for furnishing the correct data on the micro-oscillograph, and to Fred A. Koether for his help in preparing this paper.

Two Simple Bridges for Very-High-Frequency Use*

D. D. KING†, MEMBER, IRE

Summary—Two bridge circuits for impedance measurement at very-high frequency are described. The first is a hybrid junction tunable over the 100 to 500-Mc range, and capable of great sensitivity. The second is an untuned Wheatstone type for service over the same frequency range. Design features and sample performance characteristics are given for both units.

I. Introduction

BRIDGE CIRCUITS are basic in most low-frequency measurements. At microwaves, the use of waveguide bridges is also widespread. However, in the intermediate-frequency range, above 100 and below 3,000 Mc, relatively slow progress has been made in applying bridge methods. The advantages of balanced or null-type circuits are almost equally great in this portion of the spectrum, but the required electrical and mechanical design involves problems peculiar to this frequency band.

At frequencies much above 100 Mc it becomes impossible to provide the standard capacitances and resistances essential to the conventional impedance bridge. The arrangement of the bridge also presents problems due to stray capacitances and inductances. The ultimate in conventional circuitry has probably been attained by Soderman,[1] whose bridge is accurate up to a few hundred megacycles. A new null-type admittance comparator described by Thurston[2] promises to cover much of the vhf spectrum, as does a modified directional-coupler null-indicator credited to Byrne.[3] Both of these ingenious devices furnish direct readings of the unknown impedance in terms of dial settings. The former uses a wide-range multiplier, with a coaxial stub and a resistor as standards. The latter relies on broadband matched terminations.

The two devices to be considered here are bridge circuits of a simpler type. Impedance standards or their equivalent are not contained in these circuits. Instead, two coaxial terminals are available, to which arbitrary impedances may be attached. When these two impedances are equal, the bridge is balanced. The services which may be performed by such a device fall in two categories:

1. Accurately matching one impedance to another, known or unknown.
2. Detecting and measuring very small changes in an impedance.

The applications of the two bridges to be described therefore do not coincide precisely with those of the impedance measuring schemes mentioned above. The hybrid junction system is considered first.

II. A Tunable Coaxial Hybrid Junction

The hybrid junction in waveguide form is commonly used as a "magic tee" bridge. A coaxial version mentioned by Pound[4] served as the starting point for the present design. Basic to the hybrid circuit are the series and parallel connections made at the same point along a line. In order to obtain external terminals for these connections, stub supports are required. The mechanical structure devised in the present case makes use of two tunable stubs mounted in concentric fashion. The arrangement of the coaxial circuits is shown in Fig. 1.

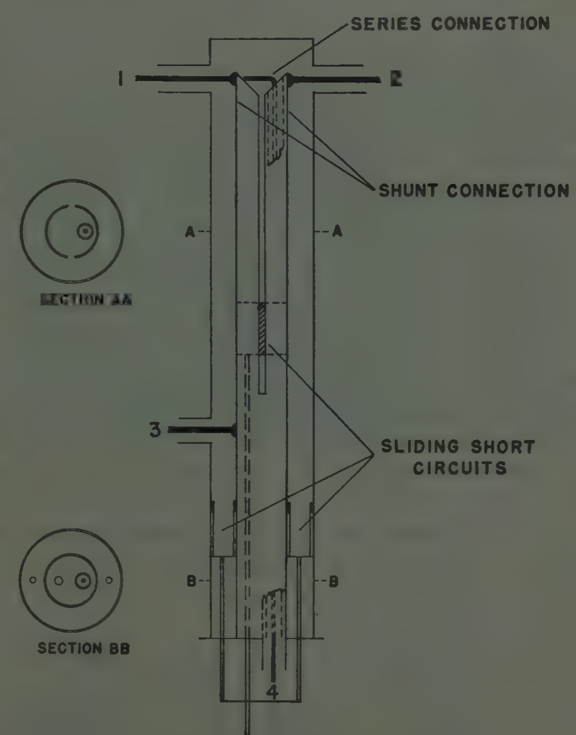


Fig. 1—Coaxial hybrid junction.

The main line has terminals 1 and 2 in the figure. Output 4 is in series with this line, and output 3 is in shunt. The outside diameter of the assembly is one inch, hence the electrical length of the actual junction point is negligible at the highest frequency used. The sliding

* Decimal classification: R244.111. Original manuscript received by the Institute May 16, 1949.

† The Johns Hopkins University, Radiation Laboratory, Baltimore, Md.

[1] Type 1601A VHF Impedance Bridge, General Radio Co., Cambridge, Mass.

[2] W. R. Thurston, "A device for admittance measurements in the 50- to 500-Mc range," presented, 1949 National IRE Convention, March 9, 1949, New York, N. Y.

[3] J. F. Byrne "A null-method for the determination of impedance in the 100–400 Mc range," *Proc. N.E.C.* (Chicago), vol. 3, pp. 603–614; 1947.

[4] R. V. Pound, "Microwave Mixers," McGraw-Hill Book Co., New York, N. Y., p. 268, 1948.

short circuit in the split inner conductor is part of an adjustable stub in parallel with the series connection 4. The plunger in the main barrel similarly is part of an adjustable stub in parallel with the shunt terminal 3. Terminals 1 and 2 are for the impedances under test; terminals 3 and 4 are for generator and detector. In the present case the generator is arbitrarily assigned to terminal 3. The equivalent circuit of the hybrid junction is shown in Fig. 2. This circuit is obtained by superposing the equivalent circuits for shunt and series tees in lines or guides. The transformer in the representation of a series tee is then replaced by an equivalent impedance Z_4. Without generator and detector connections, the bridge acts as a reactive T-network or triple-stub tuner.

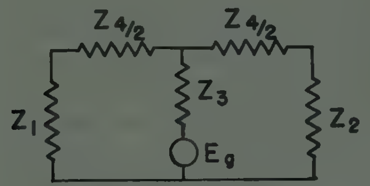


Fig. 2—Equivalent circuit of the hybrid junction.

Any residual asymmetry at the junction results in an unequal division of Z_4. This unbalance is very small when the physical symmetry is good. The shunt impedance of the two stubs mentioned above is of course included in Z_3 and Z_4 of the equivalent circuit. The theory and applications of waveguide "magic-tee" bridges is thoroughly treated in the literature,[5] and applies in large part to the coaxial version just described. A few practical aspects are mentioned here.

The simplest application of the bridge is in adjusting one impedance to precisely equal another. Exact symmetry within the bridge is then unnecessary, if an adjustable reference impedance is permanently attached to arm 1. The impedances to be equalized are then attached in turn to arm 2. The smallest change in impedance which may be detected depends in principle on the number of decibels between input power and detector noise level. The fineness of adjustment possible usually is a more practical limitation.

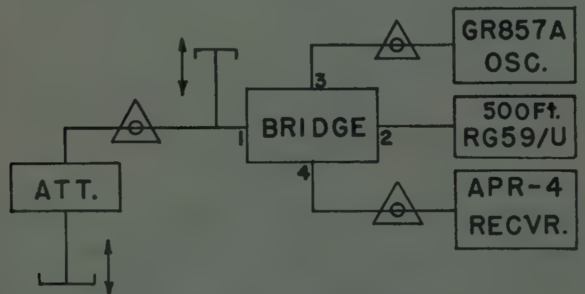


Fig. 3—Circuit for testing bridge sensitivity.

[5] C. G. Montgomery, "Technique of Microwave Measurements," McGraw-Hill Book Co., New York, N. Y., Chap. 9; 1947.

In order to test the sensitivity of the bridge, a beyond-cutoff attenuator may be used in the arrangement shown in Fig. 3. The input loop of the attenuator is a 75-Ω resistor, the output loop a copper strip. With the attenuator set at maximum attenuation, the bridge is balanced by the phase changer and stub in arm 1. The reaction of the resonant output circuit in the attenuator on the bridge null is a convenient indication of the sensitivity at balance. As an example of the sensitivity attainable, the following data were taken at 150 megacycles:

Attenuator Insertion Loss	56 db	>80 db
APR-4 Receiver Tuning Meter Reading (0–200 μamp. movement)	105 μamp.	95 μamp.

The maximum reflection coefficient at the input terminals of a matched 56-db pad, obtained by changing its output impedance, is of the order of 10^{-5}. The initial balance of the bridge is the equivalent of a perfect reference match when the pad is set at >80 db. Considered as an attenuator between power source and receiver, the bridge at balance presented an attenuation of the order of 90 db. For such critical adjustment an unmodulated signal is required to avoid the frequency modulation present in most modulated oscillators.

"Magic tee" bridges are commonly used to measure the magnitude of the reflection coefficient $|\Gamma_2|$ directly. Under these so-called "magic" conditions, $Z_1=Z_3=Z_4=Z_0$, $(\Gamma_1=\Gamma_2=\Gamma_3=0)$ where Z_0 is the characteristic impedance of the transmission lines involved. Then the input to output power ratio is[6]

$$\frac{P_4}{P_3}=\frac{|\Gamma_2|^2}{4}. \tag{1}$$

Since a shunt stub is an integral part of outputs 3 and 4, an external phase changer in these two lines suffices to provide the required matched conditions. More generally, the values of Z_3 and Z_4 may be adjusted to equal any desired reference standard Z_1. However, the sensitivity of the bridge is reduced when Z_3 and Z_4 differ appreciably from the cable impedance used. Efficient power transfer from source to bridge and from bridge to detector is thereby impaired. To adjust the bridge to the "magic" condition, a reactive load varying at an audio rate may be placed in arm 2. The other arms are then adjusted to minimize the resultant modulation in the output. A motor-driven butterfly coupled to the bridge through a pad approximates the desired performance fairly well.

Although the hybrid circuit described above is sensitive and flexible in its application, it requires careful tuning at each frequency of operation. This drawback led to the development of a second bridge of simpler design.

[6] See p. 530 of footnote reference 5.

III. A Wheatstone Bridge for VHF

The balance in a Wheatstone bridge is obtained by symmetry, and the simple bridge circuit must continue to give a null indication at any frequency at which the symmetry can be preserved. At microwave frequencies, impedances and terminals may no longer be defined in the usual way for circuits of practical dimensions. However, at frequencies in the hundreds of megacycles, definite transmission line terminals are available. Accordingly, the device shown in Fig. 4 was built to have complete physical and electrical symmetry. Any asymmetry in the structure of the detector crystal is neglected. The electrical balance is limited by the $\pm\frac{1}{2}$ per cent tolerance of the matched components.

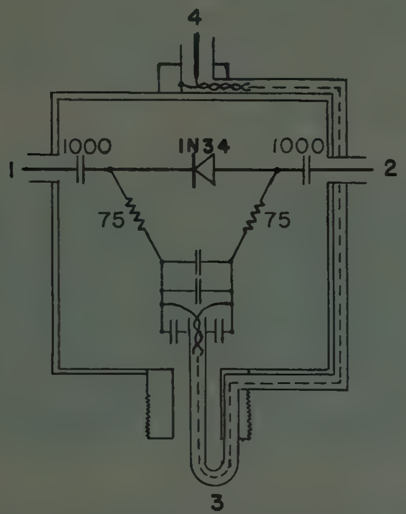


Fig. 4—Wheatstone bridge circuit. Bipass capacitors are 500 $\mu\mu$fd.

The generator terminal 3 shown in Fig. 4 is designed to mount directly in a GR type 857A oscillator. Alternatively, the loop may be mounted in one end of a beyond-cutoff attenuator. The direct mounting in the oscillator provides maximum power input over the entire 95 to 515-Mc tuning range of the oscillator. A microammeter at terminal 4 with series multiplier to protect the crystal serves as a simple null indicator. Greater sensitivity is obtained with a modulated system and tuned amplifier. In either case, the only tuning adjustment required is the oscillator dial.

The symmetry of the system is easily checked by balancing the bridge and then interchanging the loads of terminals 1 and 2. The symmetry is best at low frequencies. It is worst at high frequencies and at very low load impedances. Lack of standards prevented a quantitative check, but in the range 100 to 500 Mc the symmetry appears better than the electrical tolerances would indicate.

The sensitivity attainable is shown by Fig. 5. These data were obtained by the method shown in Fig. 3. The oscillator is, of course, direct-coupled in this case. The self-contained crystal detector feeds a tuned amplifier and Ballantine volt meter. Although the sensitivity to impedance change is much lower than that quoted for the tuned hybrid junction, it exceeds that attainable by slotted lines or directional couplers. The convenient mounting of the instrument is shown in the photograph, Fig. 6.

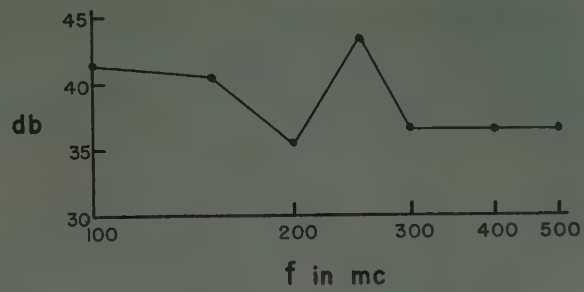


Fig. 5—Attenuator insertion loss for 1-db change in null indicator.

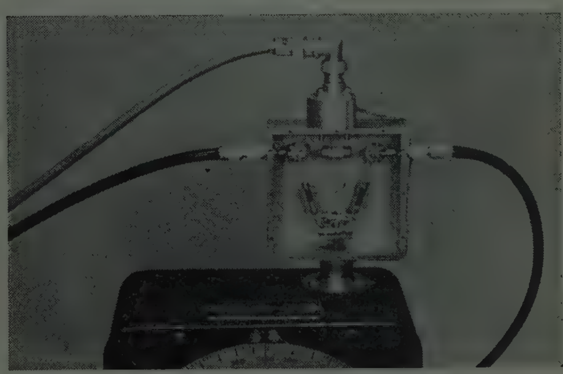


Fig. 6—Wheatstone bridge mounted on GR 857A oscillator (cover plate removed).

IV. Conclusion

Two bridge circuits useful in the 100 to 500 Mc range have been described. The first, a tuned hybrid junction, is extremely sensitive, and may be adapted to any frequency suitable for coaxial-line stubs. The second bridge is a modified Wheatstone circuit, operating without tuning adjustment over the entire 100 to 500 Mc frequency range. These circuits are applicable wherever sensitive impedance comparisons are required in the very-high-frequency spectrum.

Nonlinear Coil Generators of Short Pulses*

L. W. HUSSEY†, SENIOR MEMBER, IRE

Summary—Small permalloy coils and circuits have been developed which produce pulses well below a tenth of a microsecond in duration with repetition rates up to a few megacycles.

The construction of these coils is described. Low-power circuits suitable for different types of drive and different frequency ranges are discussed.

Introduction

NONLINEAR COIL harmonic generators have been known for some time. An approximate theory exists[1,2] which was once adequate for the frequencies and circuits used. Recently, applications of the coils as pulse or harmonic generators have been made to progressively shorter pulse durations and to higher frequencies. The most valuable recent applications are in the fundamental frequency range from 100 kc to 1 Mc or higher. Coils, to operate satisfactorily in this range, must be made physically small. An example is the little permalloy coil used in a wartime communication set.[3]

This coil served its particular purpose very well but, compared with a larger coil, it was electrically poor. The reason is obvious, if one examines the structure. The core is made of thin tape wound on a spool. Large coils made in this manner have a large amount of permalloy, plus a small percentage of spool and air inside the turns of wire. This coil has a little volume of permalloy, plus much more air and spool inside the turns of wire. This is similar to putting a sizable air core coil in series with the nonlinear one. The high permeability of unsaturated permalloy makes the air core unimportant, but when the core is saturated, instead of the inductance dropping down to the small value for the permalloy alone, it drops to that for the permalloy plus the air core. The difficulty is unavoidable, since the spool is as small and thin as practicable. It is also important because a large ratio of unsaturated to saturated inductance is essential to good pulse generation.

New Coil Construction

The next phase of this work was based upon the most direct solution of the immediate problem: Since the spool cannot be made small enough, omit it. The resulting improvement may be illustrated by comparing two typical 20-milligram coils. One, with supporting spool, has a 0.72-cm mean diameter and the total cross-section area inside the windings is roughly 65 times as great as the cross-section of the permalloy itself. The other has a mean diameter of 0.55 cm and the ratio of total cross-section area to permalloy is 8 to 1. The unsaturated permeabilities of the two coils are comparable, but the apparent saturation permeability[4] of the new coil can readily be reduced to about 18, while driving the coil on the spool as hard as it will stand, the apparent saturation permeability only drops to about 90. Fig. 1 shows sections of similar coils.

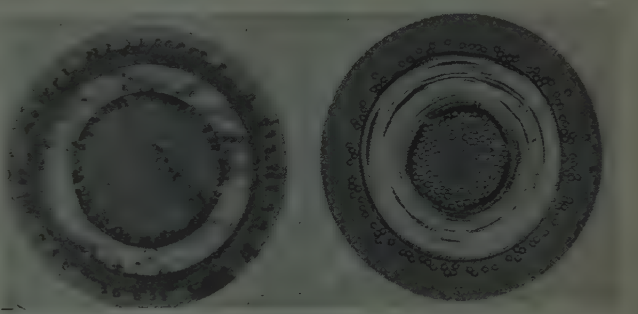


Fig. 1—A section of new coil and of coil on ceramic core showing relative volume of air and ceramic to that of permalloy tape. Ceramic coil magnified 4×. New coil 6.6×.

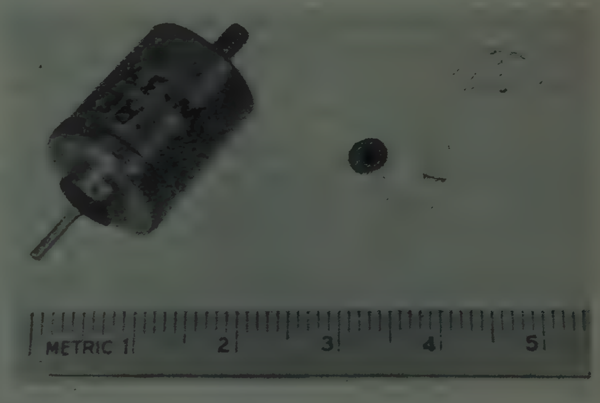


Fig. 2—Typical two-winding permalloy coil and potted single winding coil. A centimeter scale is shown for size comparison.

The new coils are units like those of Fig. 2, made of thin, insulated, molybdenum permalloy tape. They are comparable, electrically, to larger coils of the same material with the added factor that they can be used to a megacycle or higher.

A 0.25-mil tape is generally used. Thicker tape (1 mil, for example) has better low-frequency properties, but eddy current losses prohibit its use above the voice range. Thinner tapes (0.125 and 0.05 mil) are much

* Decimal classification: R355.913.3. Original manuscript received by the Institute, June 10, 1949; revised manuscript received, September 13, 1949.

† Bell Telephone Laboratories, Inc., Murray Hill, N. J.

[1] E. Peterson, J. M. Manley, and L. R. Wrathall, "Magnetic generation of a group of harmonics," *Elec. Eng.*, vol. 56, p. 995; August, 1937.

[2] E. Peterson, J. M. Manley, and L. R. Wrathall, "Magnetic generation of a group of harmonics," *Bell. Sys. Tech. Jour.*, vol. 16, pp. 437–456; October, 1937.

[3] L. R. Wrathall, "Frequency modulation by non-linear coils," *Bell. Labs. Rec.*, vol. 24, pp. 102–105; March, 1946.

[4] "Apparent saturation permeability" will be discussed later. It is roughly an average permeability in the saturation region.

poorer in over-all magnetic properties and difficult to handle. Any high-frequency advantage noted so far is insufficient to justify using them.

Elementary Pulse Generator

Before going into the use of the coils in circuits we may review briefly the behavior of one of them in the simplest type of pulse[5] generator. Fig. 9(a), shows such a circuit and Fig. 3 shows a typical *B-H* loop for the nonlinear coil *L*. In Fig. 9(a) L_0C_0 is tuned to the frequency of the generator wave to keep the generator current sinusoidal. *C* and *R* are a small capacitor and a small resistance whose magnitudes determine quite critically the form of the output. *R* represents the load impedance.

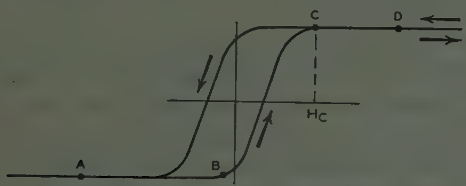


Fig. 3—*B-H* loop for permalloy coil.

Starting when the generator current is at its negative maximum (*A* of Fig. 3), the generator current increases toward zero and the coil current travels from A toward B. The small change in flux with the change in current results in small voltage across the coil and output circuit. In other words, the coil is a small inductance. Beyond B the coil suddenly becomes a large inductance. The tuning in the primary circuit prevents any abrupt modification of the generator current, so most of it is forced into the *RC* mesh, building up a large voltage. Meanwhile, the coil current changes from B to near C. Here the inductance suddenly becomes very small again, and the capacitor discharges rapidly, the high amplitude discharge current flowing in *L* and *R*. The wave form of the current pulse is determined by *RC* and the residual inductance of *L*. When *C* is discharged, the system is ready to repeat in the reverse direction.

The following characteristics of this circuit are important to all coil pulse generators:

(1) The drive should be tuned, or otherwise provide a high generator impedance to harmonics, for a large voltage build-up on *C* and an undistorted discharge pulse.

(2) The coil must behave like a switch between a very large and a very small inductance for efficient operation.

(3) The most important characteristic of the driving generator current is its rate of change while the coil is being driven from saturation in one direction to saturation in the other. Beyond that region, the large discharge current takes control. The loop of Fig. 3 is not independent of the drive speed and wave form. Eddy currents make the loop fatter, so the magnetizing force to attain saturation (C) moves out as one increases the speed of the drive.

(4) Both the amplitude and form of the discharge depend critically on the discharge condenser and resistance: For a given drive and coil the discharge condenser must be small enough to be charged up to a high voltage by the drive current, but large enough to supply the power for a high amplitude current pulse. The discharge closely resembles the free discharge of a capacitor in a simple *L*, *R*, *C* circuit. The resistance should be so chosen that the circuit is nearly critically damped. If *R* is too small, an oscillation made of repeated pulses results every cycle; if too large, a very broad, low amplitude pulse results. Figs. 4 to 7 illustrate how the pulse duration (measured at half amplitude) and amplitude vary with *R* and *C*.

Drive Circuits

As soon as attempts are made to generate short pulses (for example shorter than 10^{-7} seconds duration) the driving generator becomes a problem. The coils are small and must be driven with a very steep current wave

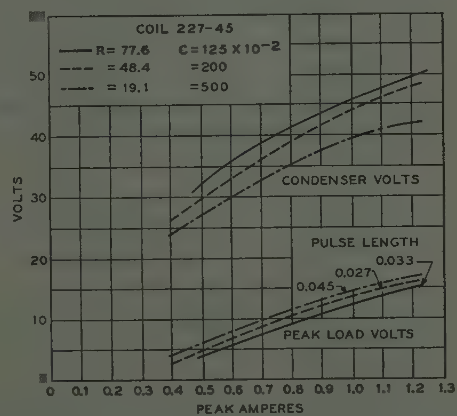


Fig. 4—Peak voltage on the discharge capacitor and peak pulse voltage on load resistance versus amplitude of sinusoidal drive current for typical cases. Pulse duration (at half amplitude) in microseconds shown at three points. Repetition rate 8.6 kc.

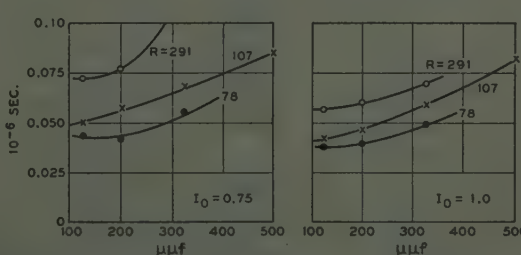


Fig. 5—Pulse duration (at half amplitude) versus discharge condenser capacity for typical sinusoidal drive current amplitudes and load resistance magnitudes. Repetition rate 8.6 kc.

[5] "Pulse generator" and "harmonic generator" of course are practically the same thing, when one is interested in a very narrow pulse or a broad band of harmonics.

front. The circuit of Fig. 9(a)—if, for example, E is an ordinary vacuum-tube amplifier and the drive low frequency—becomes very high power and inefficient.

The simplest solution, particularly starting with a pulse or square wave which must be made much narrower, is a pentode drive (Fig. 8). The pentode and choke furnish the high impedance source. This circuit, when efficient, is inherently one-sided, requiring a push-pull modification to produce equal pulses of the two polarities. The output pulse in one direction will be large while that in the opposite direction may be small or missing. In most cases this is desirable. With an unsymmetric drive, however, care should be taken that the coil current reverses sufficiently for the complete B-H loop to be traversed at each cycle.

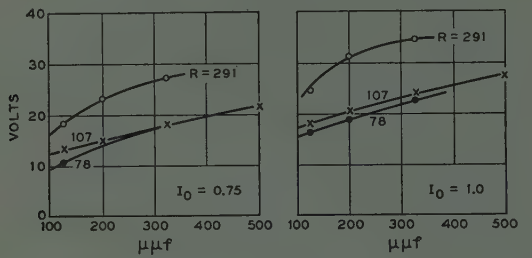


Fig. 6—Variation of peak-output pulse amplitude with discharge condenser capacity under same conditions as in Fig. 5.

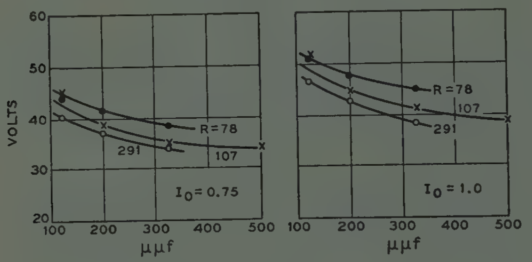


Fig. 7—Variation of peak discharge capacitor voltage with capacity under same conditions as Fig. 5.

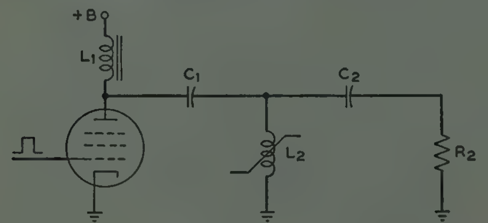


Fig. 8—Pentode drive. The sine wave generator and tuned circuit is replaced by a pentode amplifier driven, preferably, by a pulse with a steep leading edge.

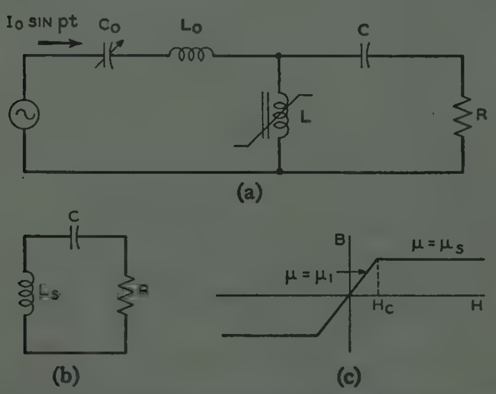


Fig. 9—Idealized harmonic generator used in analysis.

A very efficient drive, particularly if a small driving pip is available out of a high impedance generator, is a blocking oscillator.[6,7] Since the tube may be cut off during most of the cycle, very large current amplitudes can be obtained from small tubes with very small average power. The very large, short duration, cathode current drives the coil very well. Such an oscillator with a miniature tube, such as a 6AK6, and an ordinary pulse transformer, will work up to around 0.5-Mc repetition rate. The same tube, with a special transformer made with one of the small toroidal cores described herein, can be used above 1 Mc.

A tendency to jitter which seems to be inherent in blocking oscillators, is the main limitation, making this circuit unsatisfactory in some cases where precision timing is involved.

For a small sinusoidal signal from 100 kc up to 1 Mc or more as the driving source, a tuned amplifier may be used. L. F. Koerner operated a system by putting the little coil and discharge circuit in series with the inductance of the tank circuit. The large circulating current of the antiresonance drove the coil. This works well when the harmonic generator impedance is too small to greatly decrease the Q of the tank circuit. A. N. Garblik developed a circuit for relatively higher impedance harmonic generators by tapping the tank coil, making an autotransformer. A conventional tuned pulse generator circuit is then attached.

Transformers

These coils may be constructed with two or three separate windings. They may be used as pulse transformers, for pulses a fraction of a microsecond in duration. They have been used (as previously mentioned) as transformers in blocking oscillators, where higher speeds are desired than obtainable with commercial pulse transformers. If pulses from an inherently unilateral device (such as a blocking oscillator driven pulse generator) have the wrong polarity, they may be turned over by putting the discharge circuit in the secondary of a two-winding coil.

Design

As has been remarked, any computation for design purposes is difficult and approximate. Starting with elements of the correct order of magnitude, results can be obtained quite rapidly by cut-and-try methods, a good oscilloscope, and a little experience. The magnitudes involved may be obtained from the samples in the following tabulation. In Table I,

[6] F. E. Terman, "Radio Engineers' Handbook," McGraw-Hill Book Co., New York, N. Y.; p. 514, 1948.
[7] O. S. Puckle, "Time Bases," John Wiley and Sons, Inc., New York, N. Y., Chap. V, 1943.

f_0 = fundamental drive frequency
w = core weight in grams
a = core cross-section area
d = core mean diameter
n = number of turns of wire
V = peak output pulse voltage
τ = pulse duration, measured at half amplitude.

Some work has been done on design, based on that of Peterson, Manley, and Wrathall.[1] In their work, the B-H loop is approximated by broken line segments (Fig. 9(c)). The simple sinusoidal drive circuit (Fig. 9(a)) is assumed.

Their relations for inductance and peak charge on the capacitor, as a function of circuit parameters and drive current amplitude, may be used except that the maximum permeability, μ_1, and the point H_c on the saturation curve are different from their values because of thin tape and high frequency. In the present work $\mu_1 = 2{,}500$, $H_c = 1.8$ give a fair check with experiment.

When the coil saturates and the condenser starts to discharge rapidly, the circuit becomes essentially that of Fig. 9(b), where L_s is the inductance with $\mu = \mu_s$. If μ_s were actually a constant, the problem would be simple. The form of an L, R, C discharge current, or voltage across R, is given by Fig. 10, and the relation between pulse duration (length of first lobe of discharge), k, and RC given by Fig. 11. The ratio of peak output voltage, V_R, to peak capacitor voltage depends only on k (Fig. 12).

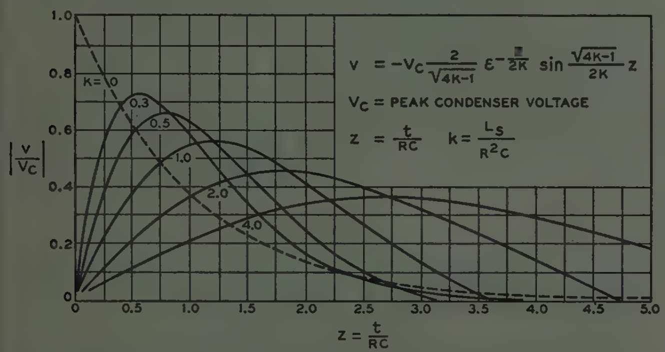


Fig. 10—Pulse voltage wave form versus discharge circuit parameter.

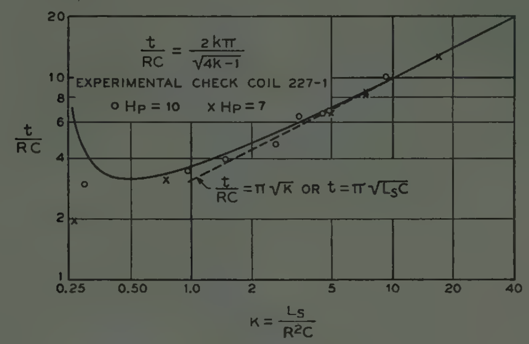


Fig. 11—Relation between pulse length (t), the time constant (RC), and the circuit parameter k.

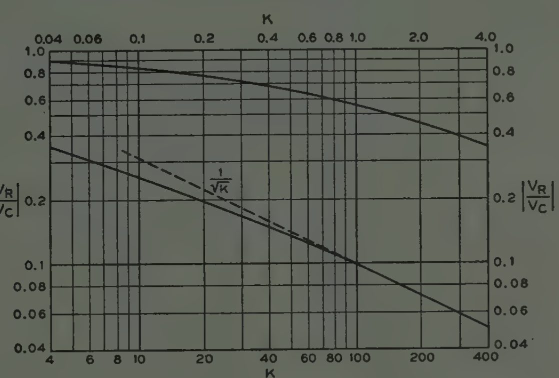


Fig. 12—Relation between ratio of peak-output pulse voltage to peak discharge capacitor voltage and the circuit parameter k.

TABLE I

Type of Drive	Coil	f_0 kc	W Grams	a cm	n	d cm	R ohms	C $\mu\mu$f	V_{peak} volts	τ μ sec	
Autotransformer (with 6AK6)	W34405	667	0.01	0.0012	90	0.3	820	50	30	0.02	(Note A)
Pentode (with 50-B-5)	236-1	672	0.003	0.0006	50	0.18	100	56	50	0.01	
Sinusoid 1 amp (peak)	227-1	8.6	0.02	0.0013	90	0.55	107	200	20	0.05	
Sinusoid 0.56 amp	230-E-3	8.6	0.04	0.0047	46	0.22	107	325	21	0.85	
Sinusoid 1.1 amp	227-45	8.6	0.018	0.0028	40	0.23	48	200	15	0.035	
Sinusoid 20 ma	230-E-10	30	0.08	0.0081	87	0.22	500	250	34	0.2	
Blocking Oscillator (with 6AC7)	228-4	5.0	0.01	0.0018	44	0.33	50	100	30	0.025	(Note B)
Autotransformer (with 6AK5)	227-31	1000	0.01	0.0016	50	0.21	100	40	44	0.025	

Note (A)—R is unusually large. It was the grid resistance of a small tube, which made the effective R smaller. V_{peak} and τ are estimated.
Note (B)—The pulse transformer used was a special one 234-E-2. W=0.08 gms. d=0.22 gm. a=0.008 cm². n=30:60.

While μ_s is not constant, it has been found empirically that one may start with the ratio of peak load voltage (V_R) to peak capacitor voltage (V_c) and determine an effective k from Fig. 12. Having k, and the circuit parameters R and C, an effective value for L_s and hence an apparent saturation permeability, μ_s is determined. The results check very well in both pulse duration and wave form (Fig. 11), except for $k<0.5$ where the pulse dies out slowly.

In this manner a range of values of μ_s has been determined which are functions of RC, k, and H_p, where H_p is the peak magnetizing force produced by the discharge current. Over the range of experiment, μ_s depended only on these parameters. Using data obtained in this manner, the curves of Fig. 13 were obtained. While these data are incomplete, they give useful information over a range for which the little coils have been found valuable. The dashed curves are simply the relationship between RC and K of Fig. 11, replotted for constant pulse duration.[8]

From these curves one can see what pulse durations are practicable and what values of H_p, μ_s, k, and RC can be used for a given pulse duration. For example, if one wants $\tau=0.05\times10^{-6}$ sec and $k=1$, $RC=0.028\times10^{-6}$ and $H_p=15$, $\mu_s=35$ or $H_p=20$, $\mu_s=27$ will do. There are, of course, a whole range of values of μ_s corresponding to H_p intermediate or greater than the values given. Since the output current is proportional to H_p, the greater H_p the greater the amplitude of the output pulse.

The factor k was chosen unity because it gives the best pulse shape. A smaller k gives a broad, low, pulse; for larger k, small secondary pulses appear. In case it is found that a sufficiently narrow pulse cannot be generated with $k=1$ and the coils and power available, a larger k may be used.

Having chosen H_p, RC, and k, values for R and C may be tried, the final choice depending on the power available and the output amplitude desired.

Some information can be obtained by extrapolation, even into the range of smaller RC and smaller pulse duration. Here data are nonexistent or too fragmentary to be used, but it is evident that, for example, for a pulse duration of 0.01 microsecond with a shape factor k of reasonable size, H_p must be 20 or greater and RC must be of the order of magnitude 5×10^{-9}.

No theory has been attempted concerning power dissipation and over-all efficiency. The maximum power that the coils will stand is about one watt, largely (particularly with sinusoidal drive) copper loss. Excepting the blocking oscillator and, under some circumstances possibly the pentode drive, the loss in the driving network is much greater than the total power put into the coil and discharge circuit.

[8] "t" of Fig. 11 is the pulse duration at the base or crossover. It is assumed that the half amplitude pulse duration is approximately $\tau=t/2$.

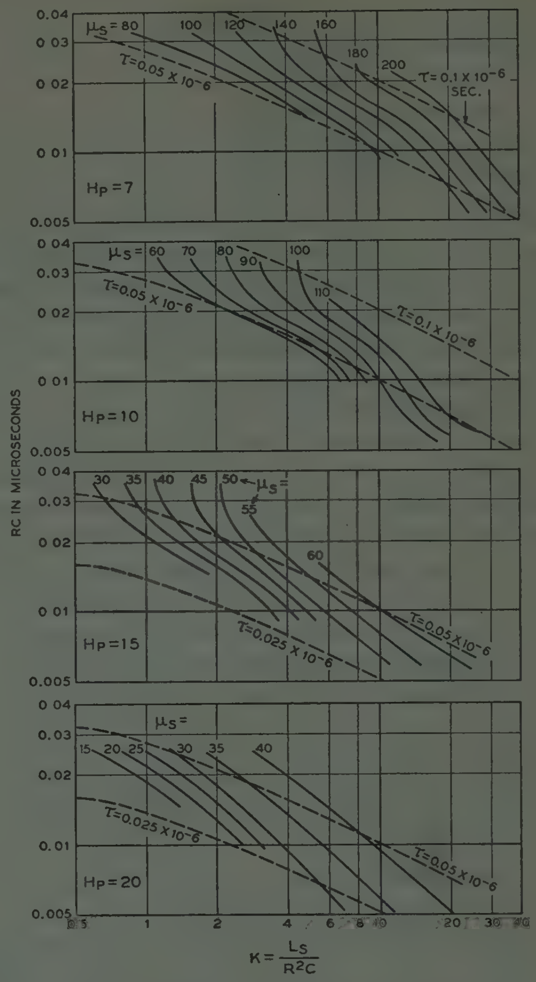


Fig. 13—Dependence of saturation permeability μ_s, on circuit parameters, k, RC, and peak magnetizing force.

A final comment may be made comparing the coil pulses with alternative devices, such as the various vacuum-tube circuits. For pulses 1 microsecond or more in duration, alternative devices (such as multivibrators and blocking oscillators) are simple and efficient, so any special advantage of the coil pulse depends on the presence of a suitable, large current power source which can be utilized efficiently. For very short pulses, vacuum-tube pulsers tend to become brute force, multitube devices and the use of the relatively simple, compact coil may mean considerable saving in space and vacuum tubes.

Acknowledgments

This work is part of an extensive program of investigation of nonlinear magnetic devices which has been carried on until recently under the direction of E. Peterson. The writer is indebted to many of his associates. Besides those already mentioned, special acknowledgment is due to A. E. Johanson, who had an active and valuable part in developing these coils.

Comparison of Measured and Calculated Microwave Signal Strengths, Phase, and Index of Refraction*

A. W. STRAITON†, MEMBER, IRE, A. H. LaGRONE†, MEMBER, IRE, AND H. W. SMITH†, ASSOCIATE MEMBER, IRE

Summary—In this paper, a comparison is made of measured signal strength-height curves and phase change-height curves for 3.2-centimeter radio waves with those determined from the corresponding measured modified index of refraction-height curves. In addition, values of the attenuation factor and modified index of refraction-height curves as determined from radio data are compared with the meteorologically measured modified index of refraction curves and the related attenuation factor. Four individual sets of radio data with path lengths of 12.3, 31.6, 40, and 47 miles are considered.

I. Radio Data

During August and September, 1947, the Electrical Engineering Research Laboratory of The University of Texas made measurements of the signal strength and rate of phase change as a function of transmitter and receiver heights. These measurements were made at a wavelength of 3.2 centimeters on an overwater path along the Gulf of Mexico near Galveston, Texas. Path lengths of 12.3, 31.6, 40, and 47.3 miles gave somewhat similar height-gain and height-phase curves. These measurements have already been reported.[1] For this paper, one set of radio data typical of each path length was chosen.

II. Meteorological Data

From metorological measurements made at the receiver site and on a ship cruising along the path, it was found that the average modified index of refraction[2] as a function of height could be represented empirically by the expression

$$M - M_0 = 0.04h + 10.5e^{-0.14h} - 10.5,$$

where M_0 is the modified index of refraction at the ground and h is the height above ground in feet. This modified index includes the earth's curvature and is defined by $M=(n+h/a-1)10^6$ where n is the actual index and a is the earth's radius. The average rather than the individual refractive index distribution was used in this study, since it was felt that the time average at a single site more nearly approached the space average, and that the average of a number of sets of readings helps to compensate for error in measurements, such as the difficulty of determining the height above the sea surface. The individual measurement of relative refractive index agreed within two M units, except in the first ten feet above the water.

III. Normal Mode Solution of Wave Equation

For distances well beyond the optical horizon, the signal strength and phase as a function of height may be obtained from the modified index of refraction distribution in terms of the sum of a number of normal modes.[3] Thus

$$\psi = e^{i\omega t - i\pi/4}\sqrt{\frac{2\pi}{kR}}\sum_m e^{-\delta_m R}U_m(h_t)U_m(h_r), \qquad (1)$$

where

R is the distance along the earth's surface
k is 2π divided by the wavelength
δ_m is the propagation function for the mth mode
$U_m(h_t)$ and $U_m(h_r)$ are the height-gain relationships as functions of transmitter and receiver height, respectively, for the mth mode.

It is convenient to express δ_m in terms of a characteristic function, λ_m, such that

$$\delta_m^2 = -k^2(1-\lambda_m). \qquad (2)$$

U_m is then given by

$$\frac{d^2U_m}{dh^2} + k^2[y(h)+\lambda_m]U_m = 0, \qquad (3)$$

where $y(h)=2(M-M_0)10^{-6}$.

From charts prepared by Pekeris and Ament,[4] and by the integral method, the value of λ for the first mode was found to be

$$\lambda = (18.2 + 0.25i)\times 10^{-6} \qquad (4)$$

and the associated attenuation with reference to the inverse square root value was 0.3 db per mile. The attenuation of the second mode was 4.2 db per mile below the inverse square root value, and the attenuation increased with increasing mode number. It then appears that, for the path lengths and the low elevations used, only the first mode need be considered.

IV. Height-Gain Function for First Mode as Determined from Meteorological Measurements

From the characteristic value for the first mode given above, it is seen that the real part of $(y+\lambda)$ be-

* Decimal classification: R271×R246. Original manuscript received by the Institute, March 31, 1949; revised manuscript received, July 15, 1949.

† University of Texas, Austin, Texas.

[1] A. W. Straiton, "Microwave phase-front measurements of 12 and 32 miles," Proc. I.R.E., vol. 37, pp. 808–813; July, 1949.

[2] J. C. Schelleng, C. R. Burrows, and E. B. Ferrell, "Ultra-short-wave propagation," Proc. I.R.E., vol. 21, pp. 427–463; March, 1933.

[3] W. H. Furry, "Theory of Characteristic Functions in Problems of Anomalous Propagation," Radiation Laboratory, Report 680; Massachusetts Institute of Technology, February, 1945.

[4] C. L. Pekeris and W. S. Ament, "Characteristic values of the first normal mode in the problem of propagation of microwaves through an atmosphere with a linear-exponential modified index of refraction," *Philosophical Mag.*, vol. 38, pp. 801–823; November, 1947.

comes zero at heights of 20.8 and 32 feet. Well below the first of these turning points, h_0^1, the magnitude of U_1 is given approximately by[3]

$$U_1 = 2C_1 e^{i\pi/4}(y+\lambda)^{-1/4} \cdot \cos\left\{k\int_h^{h_0^1}(y+\lambda)^{1/2}dh - \pi/4\right\}, \tag{5}$$

where C_1 is the normalization constant[4] given approximately by

$$C_1^{-2} = 2i\int_0^{h_0^1}(y+\lambda)^{-1/2}dh. \tag{6}$$

Well above the second turning point h_0^2, the value of U given by

$$U_1 = D(y+\lambda)^{-1/4}\epsilon^{-ik\int_{h_0^2}^{h}(y+\lambda)^{1/2}dz}. \tag{7}$$

The method used to join (6) and (7) and to establish the value of D was that of assuming that the M curve was bilinear through the turning points and of applying the equations given by Langer.[5] Although equations (5) and (7) are ordinarily applicable to strongly trapped modes, numerical substitution showed that the differential equation was approximately satisfied. It is fully realized that this solution is not exact, but it is felt that the height-gain curves thus obtained will serve as a reference to which the measured data may be compared. The height-gain curve determined from the meteorologically measured M distribution will have the same shape for all transmitter heights and for all distances for which the first mode is the only component present. However, the absolute value of the field may change with the transmitter height and with distance.

V. Phase Change-Height Function Determined from Meteorological Data

The phase change-height curve was determined from the signal strength curve by an equation given by MacFarlane.[6] The functions U and λ in (3) are expressed by

$$U = Ae^{i\phi} \quad \text{and} \quad \lambda = a + ib.$$

From this it follows that

$$\ddot{A}/A - (\dot{\phi})^2 + k^2(y+a) = 0 \tag{8}$$

and

$$2\dot{A}\dot{\phi} + \ddot{\phi}A + k^2bA = 0 \tag{9}$$

where the dots denote differentiation with respect to h. From (9) it follows that

$$\dot{\phi} = \frac{-k^2b\int_0^h A^2dh + C}{A^2}. \tag{10}$$

But $C=0$, since $\dot{\phi}=0$ when $b=0$. Then by numerical integration the value of $\dot{\phi}$ as a function of height may be obtained from the values of the magnitudes as previously determined. This curve is independent of transmitter height or path length when only one mode is present. The phase change curves thus calculated are shown later.

VI. Determination of Attenuation from Radio Measurements

MacFarlane[6] suggested the possibility of determining the index of refraction distribution with height from radio measured signal strength. Measurements at other frequencies or distances were necessary in order to determine the attenuation factor b in equations (9) and (10). This possibility was investigated by Green.[7]

However, as reported elsewhere,[5] if the rate of phase change $\dot{\phi}$ is measured simultaneously with the signal strength, simplication of the calculation results. From (9) or (10), the attenuation factor b may be determined from the measured radio data.

The method used in this study was first to draw a smooth curve through the signal strength data, which gave the value of A as a function of height. The variation of phase change as a function of height from (10) was then plotted for several assumed values of b. The measured phase change data were superimposed on these curves and the one which most nearly fitted the phase data was chosen. In fitting the points to the curves, it was usually necessary to adjust the zero phase change value. However, it is believed that this discrepancy was due to drift in the phase of the equipment after the reference angle calibration.

By this method, it was possible to determine b to within $\pm 0.05\times 10^{-6}$, which corresponds to 0.07 db per mile.

VII. Determination of Index of Refraction Profile from Radio Measurements

From (8) it is seen that the modified index of refraction is a function of the rate of phase change, the signal strength, and the second derivative of the signal strength. The value of phase change used was that obtained from the curve in the preceding section which most nearly fitted the experimental data. The first derivative of the smoothed signal strength curve was plotted against height, and this curve was in turn smoothed. The second derivative was determined from the slope of the first derivative curve. In the four exam-

[5] A. W. Straiton, "An extension of MacFarlane's method of deducing refractive index profiles from radio data," Letter to Editor, *Jour. Appl. Phys.*, vol. 20, No. 2; February, 1949.

[6] G. G. MacFarlane, "A method of deducing the refractive-index profile of a stratified atmosphere from radio observations," *Meteorological Factors in Radio Wave Propagation*, April, 1946; Report of a Conference of The Physical Society, London.

[7] J. W. Green, "Theoretical deduction of refractive index profiles from radio observations," presented, URSI/IRE Meeting, Washington, D. C., October 8, 1948.

ples shown in this paper, the refractive index curve below 20 feet was determined entirely by the signal strength curve, and above 30 feet very largely by the phase change curve. The shape of the resulting M curve below 10 feet is very critically affected by the shape of the signal strength curve.

VIII. Example A—47.3-Mile Path

The example chosen for this path length was for data taken at 1515 on 5 September 1947. The transmitter height was five feet above mean sea level. Approximately three minutes were required for taking the data.

The original signal strength data are shown by the circles in Fig. 1. The curve drawn through these circles is the curve which gave a smooth first derivative. The other curve is the height-gain curve calculated from the average meteorological data. All of the values are relative to the free-space value.

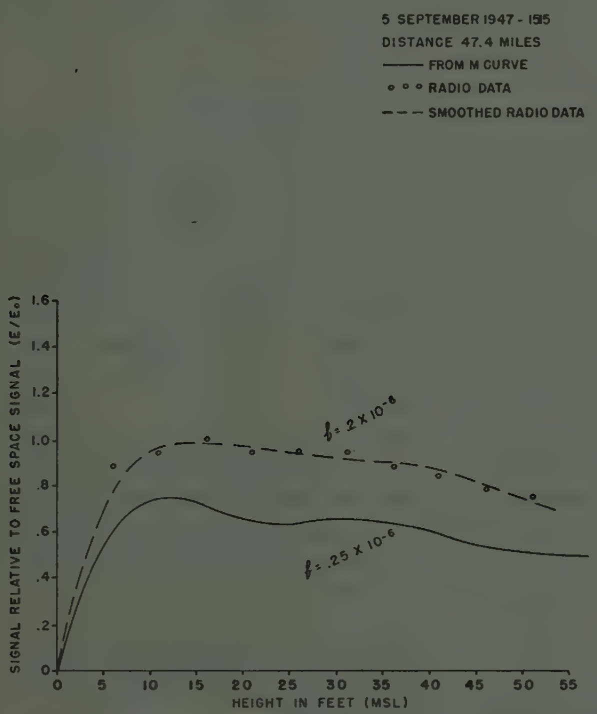


Fig. 1—Signal strength versus height for 47.4-mile path.

It is significant to note that the attenuation factor from this particular set of radio data is 0.2×10^{-6}, while that determined from the average meteorological data was 0.25×10^{-6}. This corresponds to a difference in attenuation of approximately 0.07 db per mile, or approximately 1.4 db over the path. This would account for about half of the separation of the two signal strength curves.

The phase change as a function of height for this case is plotted in Fig. 2. The circles show the measured radio data with the zero value corrected as previously described. The dotted curve is that determined from the signal strength measurements with the value of b chosen to fit the measured radio data. The solid curve represents the phase change as calculated from meteorological measurements.

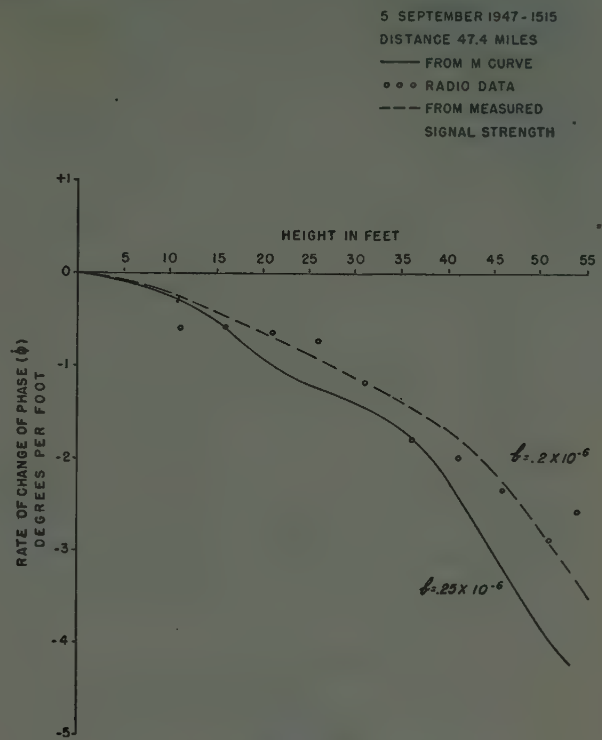


Fig. 2—Phase change-height curve for 47.4-mile path.

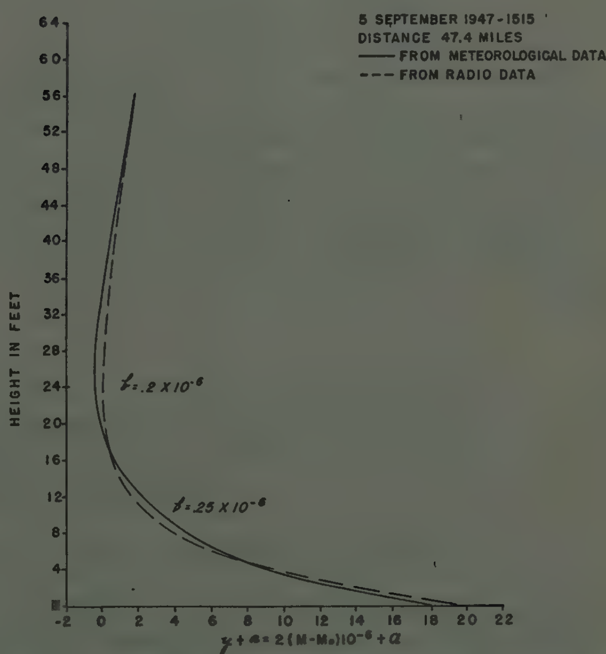


Fig. 3—M-profile for 47.4-mile data.

The modified index of refraction as a function of height for this case is plotted in Fig. 3. The solid line is the distribution determined from meteorological measurements, and the dotted line is the distribution determined from radio data. The curves agree very

closely. It should be noted that the positioning of the M curve on the $(y+a)$ base is a measure of the degree of trapping. If the curve stays to the right of the zero axis, considerable leakage takes place. If it crosses well to the left of the zero axis, the wave is "trapped." The M curve shown is seen to be a marginal case.

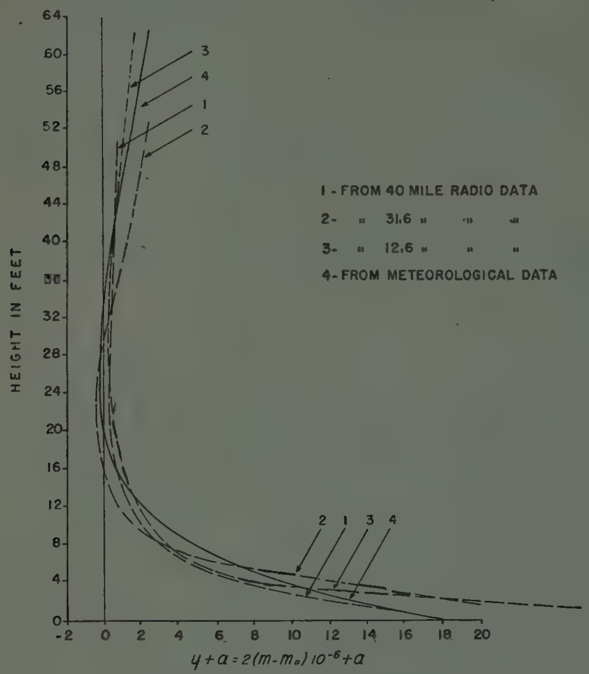


Fig. 4—Other M-profiles.

IX. Other Examples

The results of the first case and of similar studies for three other cases are shown in Table I. It is noted that the deviations of the measured signal strength curves from the ones determined from meteorological data agree in direction with that indicated by the attenuation factors. The magnitudes of the deviations shown are only approximate, since the curve shapes are not identical.

The curves of modified index of refraction from radio data for the three other cases are compared with the average measured curve in Fig. 4.

The agreement of the measured and calculated phase difference curves is not shown, but is as good or better than shown in Fig. 2.

TABLE I

Path Length Miles	*Trans. Ht. Feet (msl)*	*Date*	*Time*	*Value of $b\times10^{-6}$ Av. Met. Data*	*Radio Data*	*Deviation of Measured Curve from Theoretical Curve (db)*
47.3	5	9/ 5/47	1515	0.25	0.20	3 above
40	12	9/ 5/47	1300	0.25	0.28	1 below
31.6	10	8/28/47	1450	0.25	0.3	3 below
12.3	30	8/19/47	1612	0.25	0.3	3 below

X. Conclusion

The measured and calculated index of refraction profiles agreed to within two M units except in the first four feet. The measured and calculated signal strengths were found to have the same general shapes and to differ in magnitude by less than 4 db. The measured and calculated rates of phase change agree throughout the range considered within one degree per foot.

The data on which this work is based were not taken with these studies in mind. Considerable difficulty was encountered in both the radio and meteorological measurements due to fluctuations of the measured quantities. It would seem desirable to use longer time constants to reduce the amount of these variations. More signal strength data at low elevations and phase and signal strength at higher elevations seem desirable. Further overwater measurements are planned for the near future which will incorporate these plans.

However, in spite of these difficulties, it is felt that the attenuation and refractive index may be determined satisfactorily from radio data.

A Coupled "Coaxial" Transmission-Line Band-Pass Filter*

J. J. KARAKASH†, ASSOCIATE, IRE, AND D. E. MODE†, SENIOR MEMBER, IRE

Summary—Band-pass filters employing tuned coupled circuits are used to advantage at low frequencies and, because they have good electrical characteristics, it may be desirable to adapt them to the very-high frequencies. However, at frequencies bordering on the microwave region, it is no longer possible to treat such circuit elements in the usual low-frequency fashion, the distributed character of the parameters becoming of prime importance. This paper describes a band-pass filter in which the coupled elements consist of a shielded pair of cylindrical conductors. Because the analysis is based upon transmission-line theory, the results are valid and usable at microwave frequencies until transmission other than the *TEM* mode becomes bothersome.

* Decimal classification R386.1. Original manuscript received by the Institute, January 10, 1949; revised manuscript received, August 3, 1949. This work was done under contract W28-099-ac-128 between the Air Matériel Command, Watson Laboratories, and Lehigh University.

† Lehigh University, Bethlehem, Pa.

I. Introduction

Coupled tuned circuits have long been used as band-pass filters for they may be designed to give satisfactory performance in both the pass band and in the adjacent stop bands. An attempt to design such filters for use at microwave frequencies finds the elements involved reduced to little more than leads, and a transmission-line approach to the problem must, therefore, be made if a design is to be anything more than a "cut and try" process.

An elementary mechanical layout of the coupled transmission-line band-pass filter to be discussed is shown in Fig. 1. The tuned circuits are the transmission lines and the coupling is that existing between them. Radiation losses are eliminated due to the shield; in this sense, the system will be referred to as a coaxial filter.

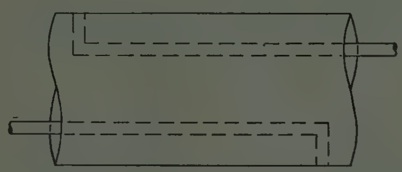


Fig. 1—Elementary prototype section of a coupled "coaxial" transmission-line band-pass filter.

II. Transmission in a Coupled "Coaxial" System

Equations for the current and voltage functions along two identical mutually coupled transmission lines such as shown in Fig. 2 have been presented by Fuchs.[1]

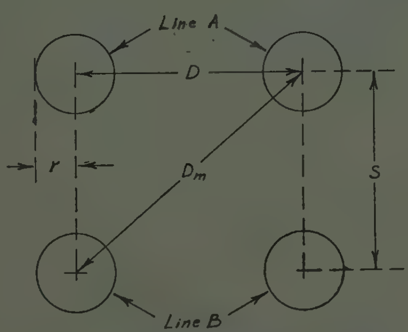


Fig. 2—A balanced coupled "coaxial" transmission line, in cross section.

These are

$$E_A = a_1\epsilon^{j\beta l} + b_1\epsilon^{-j\beta l} \tag{1}$$

$$E_B = a_2\epsilon^{j\beta l} + b_2\epsilon^{-j\beta l} \tag{2}$$

$$I_A = \frac{Z_0a_1 - Z_ma_2}{Z_0^2 - Z_m^2}\epsilon^{j\beta l} - \frac{Z_0b_1 - Z_mb_2}{Z_0^2 - Z_m^2}\epsilon^{-j\beta l} \tag{3}$$

$$I_B = \frac{Z_0a_2 - Z_ma_1}{Z_0^2 - Z_m^2}\epsilon^{j\beta l} - \frac{Z_0b_2 - Z_mb_1}{Z_0^2 - Z_m^2}\epsilon^{-j\beta l} \tag{4}$$

in which the subscripts A and B serve to identify the transmission lines, l is the line length, Z_0 is the characteristic impedance of either line alone, Z_m is the mutual impedance between lines, and a_1, b_1, a_2, and b_2 are constants. For the system shown in Fig. 2,

$$Z_0 = 120 \log_\epsilon \frac{D}{r}$$

$$Z_m = 120 \log_\epsilon \frac{D_m}{S} \tag{6}$$

neglecting proximity effects and assuming perfect conductors.

Equations (1) through (4) may be applied to the balanced coupled coaxial transmission-line system shown in cross section in Fig. 3 if Z_0 and Z_m are properly interpreted. In order to derive the required relations, McCracken[2] has utilized the bilinear transformation

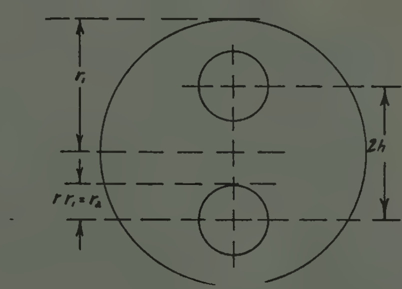


Fig. 3—Cross section of two parallel open-wire transmission lines.

$$W = \frac{aZ + b}{cZ + d} \tag{7}$$

evaluating the complex constants a, b, c, and d so that the configuration in the Z plane shown in Fig. 4(b) transforms, in the W plane, to the arrangement of Fig. 4(a). The large circle in the Z plane becomes the u axis in the W plane. Electrically, therefore, the bilinear transformation converts the dual coaxial line into an equivalent open-wire pair. Using the method of images, the conducting plane may be replaced by the two conductors shown below the u axis in Fig. 4(a), so that the system has the same geometry as that of Fig. 2. From (6) with the appropriate dimensions obtained from Figs. 2, 3, and 4(a), there results, using transformation (7),

[1] Morton Fuchs, "Intercoupled transmission lines at radio frequencies," *Elec. Commun.*, vol. 21, no. 4, pp. 248–256; 1944.

[2] L. G. McCracken, Unpublished Master's Thesis, Lehigh University, June, 1947.

$$Z_m = 120 \log_e \frac{1 + h^2 + r^2}{2h}. \tag{8}$$

The expression for Z_0 is had by calculating the capacitance C (per unit length) of two eccentric cylinders one within the other and utilizing the relations,

$$L = \frac{1}{Cc^2}$$
$$Z_0 = \sqrt{\frac{L}{C}} \tag{9}$$

where
L = the inductance of the line per unit length
c = the velocity of light.
The result is

$$Z_0 = 60 \cosh^{-1} \frac{1 - h^2 + r^2}{2r}. \tag{10}$$

The electrical performance of a coupled coaxial system (Fig. 3) is then expressible in terms of (1) through (4), with Z_m and Z_0 coming from (8) and (10).

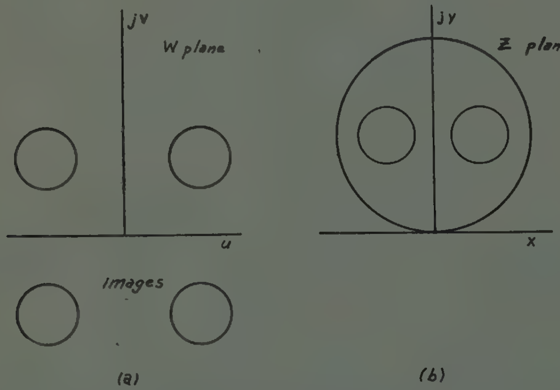


Fig. 4—Illustrating the complex transformation from a coaxial to an open-wire system.

III. General Properties of the Coupled Transmission-Line Filter

Calling the electrical length of the filter $\theta = \beta l$, using the defining equations (11) and (12),

$$E_s = AE_L + BI_L \tag{11}$$
$$I_s = CE_L + DI_L \tag{12}$$

the parameters A, B, C, and D for conditions shown in Fig. 5 are,

$$A = -\frac{1}{K} \cos\theta \tag{13}$$

$$B = -j \frac{Z_0}{K}(1 - K^2) \sin\theta \tag{14}$$

$$C = \frac{j \csc\theta(\cos^2\theta - K^2)}{Z_0(1 - K^2)K} \tag{15}$$

$$D = A \tag{16}$$

in which

$$K = \frac{Z_m}{Z_0}. \tag{17}$$

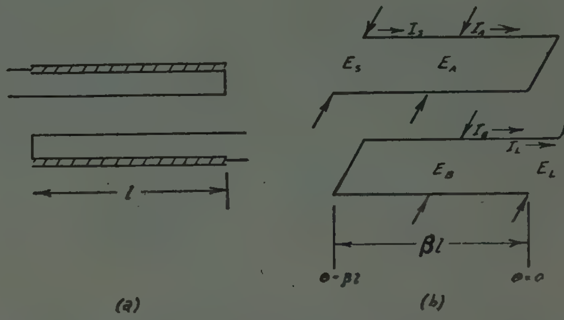


Fig. 5—A typical "coaxial" coupled transmission-line filter section. A cross section is shown in (a) and an electrical equivalent in (b).

The characteristics of the filter may be summarized as follows:

1. The pass band (defined for real Z_I) occurs for

$$-1 \leqq A = -\frac{1}{K} \cos\theta \leqq 1 \tag{18}$$

giving the cutoff conditions,

$$\theta = \cos^{-1} \pm K \tag{19}$$

from which it is seen that the pass-band center frequency f_0 occurs where the lines are one-quarter wavelength long, the width of the pass band depending upon the parameter K.

2. The reactive attenuation in the stop bands is, in nepers,

$$\alpha = \cosh^{-1} \left| \frac{\cos\theta}{K} \right|. \tag{20}$$

3. The normalized image impedance of the filter is given by,

$$\frac{Z_I}{Z_0} = (1 - K^2) \sqrt{\frac{\sin^2\theta}{K^2 - \cos^2\theta}}. \tag{21}$$

At the midband frequency f_0 (an important point from the standpoint of impedance matching), this is,

$$\left.\frac{Z_I}{Z_0}\right]_{f_0} = \frac{1 - K^2}{K} \tag{22}$$

which reduces to unity for $K = 0.619$.

The general form of the attenuation and image-impedance functions is shown in Figs. 6 and 7, respec-

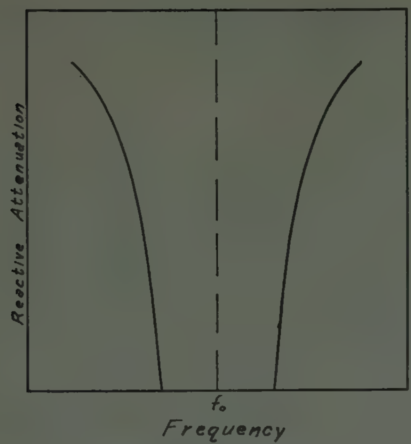


Fig. 6—Typical reactive attenuation of a coupled "coaxial" transmission line filter.

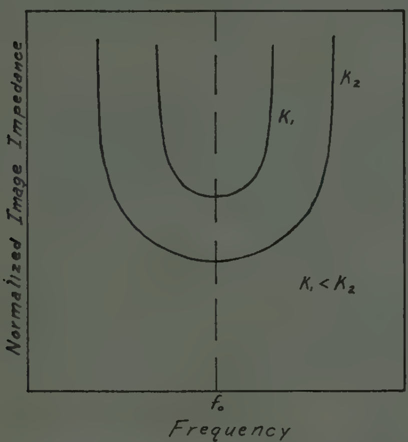


Fig. 7—Image-impedance behavior of a coupled "coaxial" transmission-line filter, K as parameter.

tively. Certain similarities in the electrical performance of the coupled coaxial transmission-line filter and its lumped-element predecessor are worthwhile pointing out. An equivalent circuit of a lumped-element coupled circuit filter is shown in Fig. 8. Assuming lossless circuit

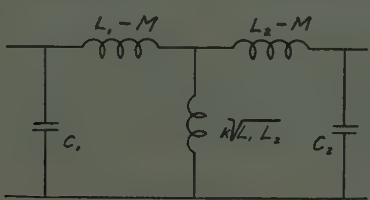


Fig. 8—An equivalent circuit of a lumped-element coupled circuit band-pass filter.

elements, that the coupled circuits are tuned to the same frequency f_0, and that $L = L_1 = L_2$, the image impedance of this filter (normalized to $Z = \sqrt{L/C}$) is

$$\frac{Z_I}{Z} = \sqrt{\frac{(1-k^2)(f_0/f)^2}{\left[(1-k) - \frac{f_0^2}{f^2}\right]\left[\frac{f_0^2}{f^2} - (1+k)\right]}} \qquad (23)$$

a typical plot of which is shown in Fig. 9. The bandwidth (defined for real Z_I) may be written,

$$BW = \frac{\omega_0}{\sqrt{1-k}} - \frac{\omega_0}{\sqrt{1+k}}. \qquad (24)$$

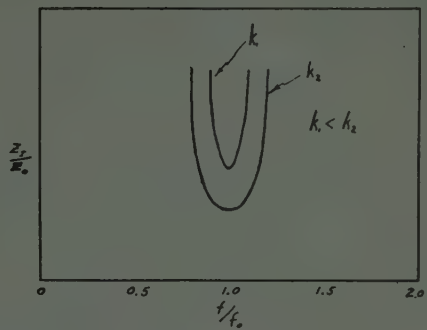


Fig. 9—Image-impedance characteristics of the lumped-element filter shown in Fig. 8.

Comparison with the coupled transmission-line filter shows that the image-impedance curves have similar form and the bandwidth is dependent upon k or K in a similar manner, becoming zero when $k = K = 0$.

IV. The Reduction of Reflection Losses

In order that the insertion loss in the filter pass band be small, the filter image impedance should match that of the circuit in which it is to work. Equation (21) shows that there are three parameters in the image-impedance equation, θ, K, and Z_0. Only two of these are independent, however, and these are required for fixing the bandwidth and center frequency of the filter. Another parameter must be introduced, therefore, so that the nominal impedance of the filter be adjustable to an optimum value. The familiar quarter-wave transformer matching scheme is therefore proposed.

Suppose that a coupled coaxial transmission-line filter has been designed so that its center frequency and bandwidth are the required ones. Such a filter is sketched in Fig. 10(a). Terminating sections are then added as

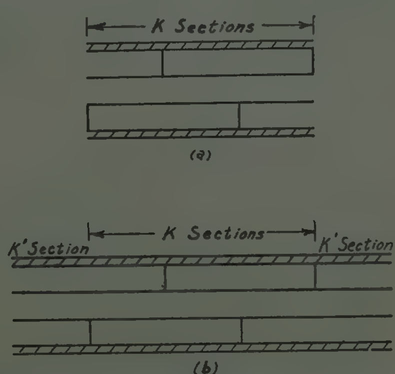


Fig. 10—The coupled "coaxial" transmission-line filter shown in cross section. (a) indicates typical midsections, and (b) shows a filter including terminating K' sections.

shown in Fig. 10(b), the K parameter (referred to as K') of these sections being chosen so that at $f = f_0$

$$Z_I' = \sqrt{Z_I Z_c} \tag{25}$$

where the prime indicates the image impedance of the terminating section, and Z_c is the external circuit impedance. If $Z_c = Z_0$, (a limiting case)

$$\frac{1 - K'^2}{K'} = \sqrt{\frac{1 - K^2}{K}}. \tag{26}$$

Two methods for the mechanical design of terminating sections are available, the first employing inner conductors of different diameter and the second a change in the eccentricity. Mechanically, the latter method is more difficult to achieve than the former. For practical narrow-band filters, $K < 0.619$ and therefore K' will be larger. For K equal to 0.619, no impedance transformation is necessary assuming $Z_c = Z_0$.

The effect of the terminating sections upon the overall image-impedance characteristic is important. A typical characteristic is shown in Fig. 11 for a single K section with two K' terminating sections.

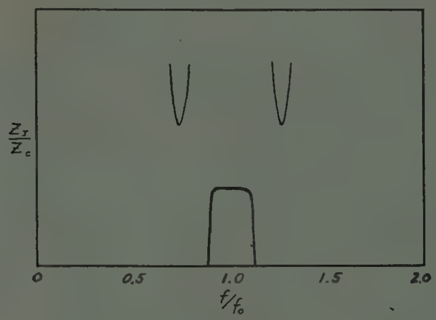


Fig. 11—Normalized image-impedance behavior of a coupled transmission-line filter having one K section together with terminating K' sections.

V. Experimental Results

As an experimental check on the theory presented, two filter models have been built and tested. The first of these had a midband frequency of 2,000 Mc, with a K of 0.44 and a K' of 0.55. The cutoff frequencies were calculated to be 2,590 and 1,410 Mc. A curve of the insertion loss *versus* frequency for this filter is shown in Fig. 12.

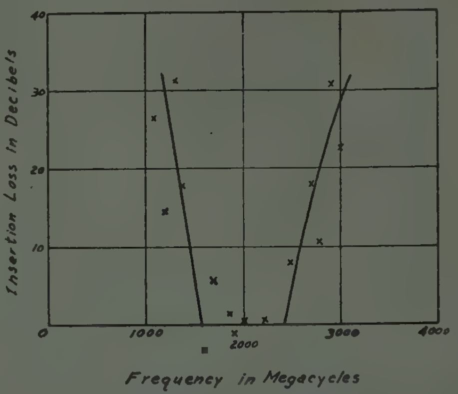


Fig. 12—Experimental insertion loss curve for a coupled transmission-line filter having parameters $f_0 = 2{,}000$ Mc, $K = 0.44$, and $K' = 0.55$.

A second filter model having a K value of 0.31 and no terminating sections was constructed for a center frequency of 1,000 Mc and a pass band between 800 and 1,200 Mc. The experimental insertion loss curve for this filter is shown in Fig. 13, where it may be seen that the cutoff frequencies tend to agree, but that the pass-band insertion loss is rather high. Without the terminating sections, the impedance match is rather poor, and this may explain the loss peaks in the pass band.

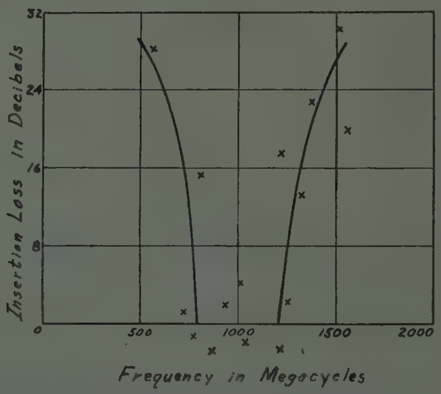


Fig. 13—Experimental insertion loss curve for a coupled transmission line filter having parameters $f_0 = 1{,}000$ Mc, $K = 0.31$.

Wide-Angle Metal-Plate Optics*

JOHN RUZE†, MEMBER, IRE

Summary—The design equations for constrained metal-plate lenses are derived. These lenses prove to have exceptional wide-angle scanning properties. The scanning aberrations are investigated by means of a power series expansion of the phase error. This analysis indicates that the square-law error is a function of lens depth and may be eliminated by the proper choice of lens thickness. It is also indicated that coma is associated with the inner lens contour, and that elliptical or double-correction-point lenses have minimum coma. The mechanism of refocussing is investigated and its first-order effect is given as a power series.

The scanning performance of various lens types is expressed by means of formulas giving the number of beamwidths scanned as a function of the f/D ratio. Experimental data are presented on a lens capable of sweeping a one-degree beam 100 beamwidths. Further data indicate the importance of smooth contours and compact structures in achieving low spurious radiation and large scanning angles.

I. Introduction

In MICROWAVE metal-plate optics the refractive property of the medium is due to the fact that the phase velocity of electromagnetic waves, confined between conducting plates parallel to the electric vector, differs from that in free space. Metal-plate lenses described heretofore in the literature[1,2] have been of a type which obey Snell's law. Here the direction of ray travel is governed by this optical relationship involving the refractive index of the medium. Such lenses we call here "normal" lenses and Fig. 1(a) shows a normal cylindrical lens. It should be noted that the metal plates are shaped to a lens contour and that the wave-front rectification is accomplished in the plane of the conducting plates (i.e., the plane of the electric vector).

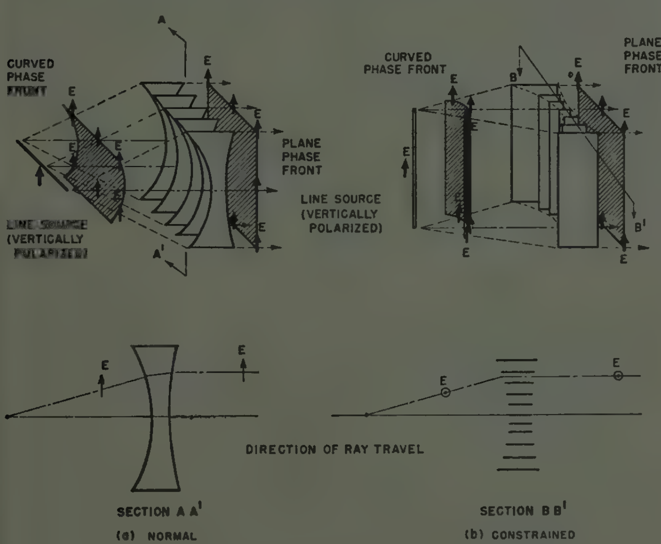


Fig. 1—Normal and constrained lenses.

The lenses described in this report (Fig. 1(b)) are of the "constrained" type. Here the rays are "guided" or "constrained" by the metal plates, and their direction is not affected by the refractive index. Obviously such lenses do not obey Snell's Law. It should be further noted that here the metal plates are merely rectangular sheets, and that the focussing is obtained normal to the constraining plates (i.e., normal to the electric vector). Constrained lenses readily lend themselves to wide-angle analysis, since the path lengths within the lens are fixed and are not a function of the angular feed position.[3]

II. Discussion

A. Derivation of Design Equations

Consider the configuration of Fig. 2 and let us inquire what conditions must be imposed upon the lens parameters so that a plane wave will be radiated at an angle α, when a source is placed at the point O.

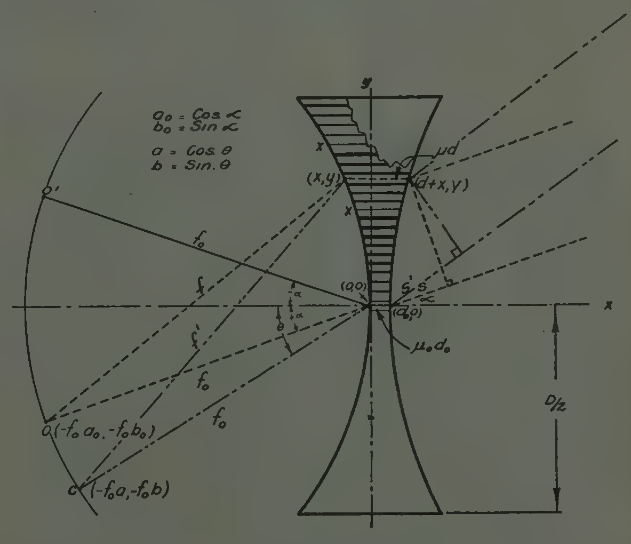


Fig. 2—Lens cross section.

For the desired plane wave, the condition which we must impose is that the phase or electrical path-length difference between a general ray and the central ray be zero or an integral multiple of a wavelength. With the aid of Fig. 2 this may be expressed mathematically:

* Decimal classification: R310×535. Original manuscript received by the Institute, January 28, 1949; revised manuscript received, September 28, 1949. Presented, 1949 National IRE Convention, New York, N. Y., March 8, 1949.

† Air Force Cambridge Research Laboratory, Cambridge, Mass.

[1] W. E. Kock, "Metal-lens antennas," Proc. I.R.E., vol. 34, pp. 828–836; November, 1946.

[2] W. E. Kock, "Metallic delay lenses," *Bell. Sys. Tech. Jour.*, vol. 27, pp. 58–82; January, 1948.

[3] For recent literature on "constrained" or "path-length" lenses, see H. B. Devore and H. Iams "Microwave optics between parallel conducting sheets," *RCA Rev.*, vol. 9, p. 721; December, 1948. And W. E. Kock, "Path-length microwave lenses," Proc. I.R.E., vol. 37, pp. 852–855; August, 1949.

$$f + \mu d - f_0 - \mu_0 d_0 - s = m\lambda \quad (1)$$

$$f = [(x + a_0 f_0)^2 + (y + b_0 f_0)^2]^{1/2}$$
$$s = a_0(d - d_0 + x) + y b_0$$
$$m = \text{any integer.}$$

Upon substitution and manipulation, we have

$$\{m\lambda - [(\mu - a_0)d - (\mu_0 - a_0)d_0]\}^2 + 2(f_0 + a_0 x + b_0 y)\{m\lambda - [(\mu - a_0)d - (\mu_0 - a_0)d_0]\} = b_0^2 x^2 + a_0^2 y^2 - 2a_0 b_0 xy. \quad (2)$$

It is to be noted that the lens parameters afford three independent variables: namely, the shape of the inner contour, the thickness of the lens, and the variation of the refractive index. We must specify three independent conditions to determine our lens uniquely. Equation (2) represents one such condition. If we limit our discussion to symmetrical structures, we require another point of perfect focus at the symmetrical point O'. This condition, obtained from (2) by using $-\alpha$ instead of $+\alpha$, yields the equation:

$$\{m\lambda - [(\mu - a_0)d - (\mu_0 - a_0)d_0]\}^2 + 2(f_0 + a_0 x - b_0 y)\{m\lambda - [(\mu - a_0)d - (\mu_0 - a_0)d_0]\} = b_0^2 x^2 + a_0^2 y^2 + 2a_0 b_0 xy. \quad (3)$$

The conditions expressed by (2) and (3) may be written more conveniently by subtracting and adding these equations and substituting, thus:

$$m\lambda - [(\mu - a_0)d - (\mu_0 - a_0)d_0] = -a_0 x. \quad (4)$$

$$x^2 + a_0^2 y^2 + 2a_0 f_0 x = 0. \quad (5)$$

For given correction points, (5) specifies the locus of the inner lens contour. This locus is an ellipse, with foci at the points O and O'. Having chosen these points, we restrict our discussion to lenses of this elliptical inner face. We are still permitted to specify a further condition; this condition, however, is restricted to the lens thickness or to the nature of the refractive index. Lenses of different mechanical and electrical characteristics will result, depending on the choice of this third condition. Conditions which suggest themselves for investigation are (1) constant refractive index, (2) shape of the outer contour, and (3) third correction point.

B. Scanning Aberrations[4]

In radar applications it is desired to scan a narrow beam over a wide angular region. With the feed at the points O and O', the diffraction or radiation pattern is that of a uniform phase aperture whose amplitude function is largely determined by the directivity of the horn illuminating the inner lens contour. At some point C, located at the general angular position θ, the phase front will deviate from the plane value. This phase deviation may be expressed as a power series across the lens aperture; that is, as

$$\delta = ay + by^2 + cy^3 + dy^4 + \cdots, \quad (6)$$

where the constants are functions only of the feed position and the parameters of the lens. These constants must be zero at the correction points or points of perfect focus.

The nature of this series and the magnitude of the various terms will determine the type of distortion of the diffraction pattern. The first term of the series represents a linear phase variation, and therefore a tilting without distortion of the radiation. The second-order term yields a radiation pattern equivalent to the "close-in" pattern of an aperture with uniform phase. This type of distortion may be eliminated by refocusing the feed. The cubic term yields a third-order aberration known as "coma." Higher-order terms exist and their aggregate forms a significant contribution to the total phase error, especially at large angles. Such terms, however, will not be considered, since individually they are smaller than the coma term; moreover, except at large angles or in cases where the coma is diminished by some means, they will not be the limiting factor on our maximum scanning angle. When such terms are of importance, the phase deviation will be determined directly without recourse to series expansion.

Let us now derive the phase deviation in the form of a power series for the constrained lens, subject to the conditions of equations (4) and (5), and to a yet unimposed third condition. The phase error at the general angular feed position θ may be written

$$\delta = f' - f_0 + \mu d - \mu_0 d_0 - s'. \quad (7)$$

Substituting (3), after some manipulation, we find,

$$\delta = f_0[1 + 2by/f_0 + b_0^2 y^2/f_0^2 + 2(a - a_0)x/f_0]^{1/2} - (a - a_0)(x + d - d_0) - yb - f_0. \quad (8)$$

When the magnitude of the sum of the last three terms of the radical is less than unity, it may be expanded as a convergent power series. It may be shown that for an f/D greater than 0.8, this expansion is permissible for all angles. We have for the first four terms

$$\delta = \frac{y^2}{2f_0}(b_0^2 - b^2) - (a - a_0)(d - d_0) + \frac{by^3}{2f_0^2}[(b^2 - b_0^2) + a_0(a - a_0)] + \frac{y^4}{8f_0^3}\{(5b^2 - b_0^2)[(b_0^2 - b^2) + a_0(a_0 - a)] + aa_0(a - a_0)^2\}. \quad (9)$$

The term involving the lens thickness may contribute both second- and higher-even-order terms. We may write for small angles

[4] An excellent reference to this material is H. T. Friis and W. D. Lewis, "Radar antennas," *Bell Sys. Tech. Jour.* vol. 26, pp. 219–247, April, 1947. The conclusions reached in the present report will be more obvious if the reader is thoroughly familiar with this reference.

$$\delta \cong \frac{1}{2}(\alpha^2 - \theta^2)\left\{\frac{y^2}{f_0} - (d - d_0)\right\} + \frac{\theta y^3}{4f_0^2}(\theta^2 - \alpha^2). \quad (10)$$

Several things should be noted about this phase-error expansion, namely:

(1) There is no "tilt" error or linear term in y.

(2) The coma term on the axis is zero. This is to be expected, since coma is an unsymmetrical error.

(3) The second-order aberration may be eliminated by choosing d properly. This type of lens may be of importance in systems where the feed is constrained to move on the focal arc. We may also conclude that the magnitude of the second-order error is a function of lens thickness or of the outer contour.

(4) The coma error is independent of how we choose the third condition. We conclude that the coma is associated with the inner lens contour, and that all symmetrical lenses with two given correction points have the same coma.

Before we consider the characteristics of particular designs, it is desirable to analyze the mechanism of refocussing.

C. Effect of Refocussing

Experimentally, refocussing consists of moving the feed radially to a point where best patterns are obtained. We can take the locus of zero refocussing as the focal arc. This definition is convenient, since some scanning systems use this arc as the natural motion of the feed.

Consider Fig. 3 where the feed has been moved inward an amount ϵf_0. The phase correction introduced by this movement is:

$$\Delta\delta = f' - f'' - \epsilon f_0. \quad (11)$$

Substituting the mathematical form for these distances and the locus of the inner contour (5), we obtain after expanding, subject to the same restrictions as previously

$$\Delta\delta \cong -\frac{\epsilon}{2}a^2\frac{y^2}{f_0} + \epsilon ab\left(a - \frac{a_0}{2}\right)\frac{y^3}{f_0^2} + \cdots, \quad (12)$$

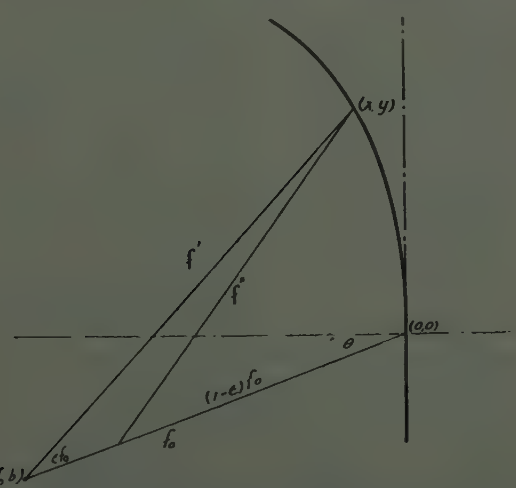


Fig. 3—The effect of refocussing.

and for small angles

$$\Delta\delta \cong -\frac{\epsilon}{2}\frac{y^2}{f_0} + \frac{\epsilon}{2}\theta\frac{y^3}{f_0^2}. \quad (13)$$

It should be noted that the effect of moving the feed from the focal arc is to introduce a second-, a smaller third- and higher-order terms. The odd-order terms are present only when the feed is off axis.

D. Investigation of Particular Cases

We are now in a position to investigate specific cases. Equations (4) and (5) (with another to be chosen at will) will determine the lens parameters uniquely. Equation (9) indicates the amount of phase distortion, and (13) the correction possible by refocussing. Various cases of interest can best be compared by specifying the third condition and tabulating the various electrical characteristics. This has been done in Table I. Fig. 4 shows the

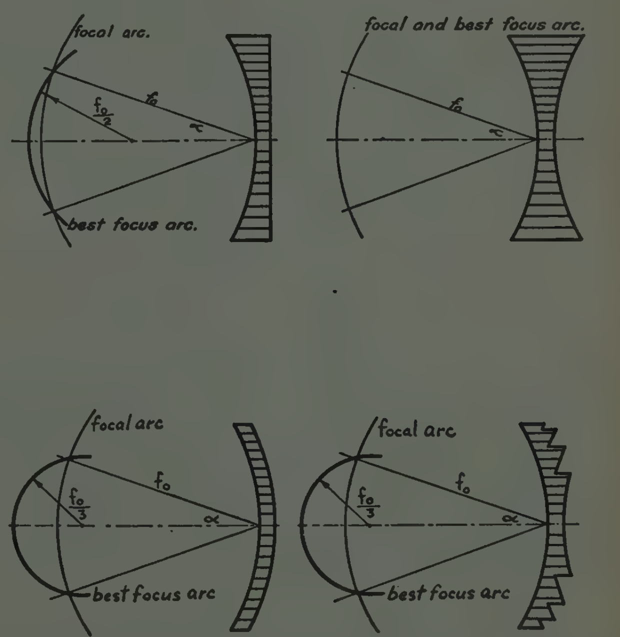


Fig. 4—Different types of lenses. (a) Straight front face; (b) no-second-order (triple correction); (c) constant thickness; and (d) constant refractive index.

physical form of these lenses. Several items should be noted from Table I, namely:

(1) The "constant thickness" lens may be attractive because of its mechanical compactness. However, it has a large second-order error when the feed is on the focal arc and requires excessive refocussing.

(2) The "straight front" lens should yield excellent patterns since refocussing to eliminate the square-law error automatically almost eliminates the third. Therefore, if refocussing is not objectionable, this lens is highly coma-corrected.

(3) To obtain the relationships for the "no-second-order" lens, we set the square-law term in (9) equal to zero. Unfortunately this condition cannot be imposed on a scanning antenna, as it is a function of the variable

TABLE I

SUMMARY OF RESULTS ON DOUBLE CORRECTION CONSTRAINED LENSES

$$x^2+a_0^2y^2+2a_0f_0x=0;\quad x=-a_0f_0\left[1-\sqrt{1-\frac{y^2}{f_0^2}}\right]\cong\frac{-a_0y^2}{2f_0}$$

Type	Third Condition	$d=g_2(y)$	$\mu=g_3(y)$	Phase Deviation		Amount of Refocussing, ϵf_0
				No Refocussing	With Refocussing	
Constant Thickness	$d=d_0$	$d=d_0$	$\mu=\mu_0+\frac{a_0x+m\lambda}{d_0}$	$\frac{y^2}{2f_0}(\alpha^2-\theta^2)-\frac{\theta y^3}{4f_0^2}(\alpha^2-\theta^2)$	$\frac{\theta y^3}{4f_0^2}(\alpha^2-\theta^2)$	$(\alpha^2-\theta^2)f_0$
Straight Front Face	$d=d_0-x$	$d=d_0-x$	$\mu=\mu_0\frac{d_0}{d}+\frac{m\lambda}{d}$	$\frac{y^2}{4f_0}(\alpha^2-\theta^2)-\frac{\theta y^3}{4f_0^2}(\alpha^2-\theta^2)$	Negligible	$\frac{1}{2}(\alpha^2-\theta^2)f_0$
No Second Order Error	$d=d_0+\frac{y^2b_0^2}{2f_0(1-a_0)}$	$d=d_0+\frac{y^2b_0^2}{2f_0(1-a_0)}$	$\mu=\mu_0\frac{d_0}{d}+\frac{m\lambda}{d}+\frac{a_0}{d}(d-d_0+x)$	$-\frac{\theta y^3}{4f_0^2}(\alpha^2-\theta^2)$	$-\frac{\theta y^3}{4f_0^2}(\alpha^2-\theta^2)$	Negligible
Third Correction Point	Perfect Focus on Axis	$d=d_0-x-\frac{f_0-\sqrt{(f_0+x)^2+y^2}}{1-a_0}$	$\mu=\mu_0\frac{d_0}{d}+\frac{m\lambda}{d}+\frac{a_0}{d}(d-d_0+x)$	$-\frac{\theta y^3}{4f_0^2}(\alpha^2-\theta^2)$	$-\frac{\theta y^3}{4f_0^2}(\alpha^2-\theta^2)$	Negligible
Constant Refractive Index	$\mu=\mu$	$d=d_0+\frac{m\lambda+a_0x}{\mu_0-a_0}$	$\mu=\mu_0$	$\frac{y^2}{2f_0}(\alpha^2-\theta^2)-\frac{\theta y^3}{4f_0^2}(\alpha^2-\theta^2)+\frac{1}{2}\frac{(\alpha^2-\theta^2)}{\mu_0-a_0}\left[\frac{a_0^2y^2}{2f_0}-m\lambda\right]$	$\frac{\theta y^3}{4f^2}(\alpha^2-\theta^2)+\frac{1}{2}\frac{(\alpha^2-\theta^2)}{\mu_0-a_0}\left[\frac{a_0^2y^2}{2f_0}-m\lambda\right]$	$(\alpha^2-\theta^2)f_0$

feed position. However, we can eliminate the second-order error on axis, where it is greatest. Such a lens would be useful in systems where the feed is constrained to move in the focal arc, and little or no refocussing is permitted. The "triple-correction point" lens is very similar to the "no-second-order" lens, and its equations could be obtained by writing (3) for perfect correction at the axial point.

(4) The "constant refractive index" lens has manufacturing advantages in that a constant plate spacing is used. It may be reduced in thickness by stepping; that is, by using integer values of m in the thickness formula. When this is done the phase deviation has a discontinuous term which is approximately zero at the beginning of each step. Fig. 5 indicates this phase variation across the lens aperture. The discontinuous term is independent of the f/D ratio of the system, and therefore sets an upper limit to the scanning angle of this type of lens. This term may assume large proportions if the refractive index or the correction angles are unhappily chosen. If the lens were not stepped, this term would reduce the normal second-order error; in fact, the refractive index could be chosen to eliminate completely this type of error. Stepping is necessary because of tolerance considerations in constant-refractive-index lenses; however, considerable second-order correction can be achieved by judiciously placing or eliminating the center steps.

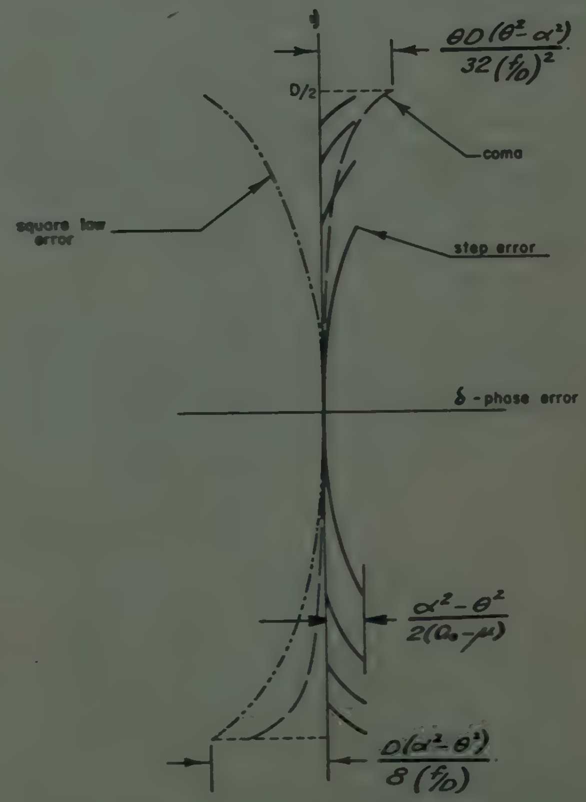


Fig. 5—Phase error for constant refractive index lens.

E. Nonelliptical Lenses

So far, we have investigated lenses with two symmetrically located correction points. The constraints of these two points have used two of the three available conditions. Such lenses, perforce, have an elliptical inner contour and an invariant coma. Let us now make available for our use two conditions by specifying only a single correction point.

It can be shown[5] that lenses with two correction points have minimum coma and that, therefore, the nonelliptical cases have little interest for us as far as wide-angle applications are concerned.

III. Design Considerations

In the previous sections we have derived the equations governing the lens parameters, and by means of a power series expansion of the phase errors have obtained a means of estimating the radiation performance. This section is devoted to some engineering design considerations.

A. Maximum Scanning Angle

To determine the maximum scanning angle of the various lens types we must set some limit to the phase deviation beyond which the radiation patterns become unacceptable. Footnote reference 4 aids in forming such an estimate. In this report the following criteria for the various types of aberrations have been chosen:

(1) Step or discontinuous error $\lambda/8$
(2) Second-order error $\lambda/6$
(3) Coma error $\lambda/4$.

It must be admitted that these criteria are rather arbitrary, depending not only on the stringency of the pattern requirement, but also on the illumination taper employed.

Normally the inner contour of the lens is excited by an electromagnetic horn. The directivity of this horn determines the illumination. The greater the illumination taper, the lower the side lobes, the wider the main beam, and the lower the gain.[4] However, if the energy which is lost over the antenna edge because of insufficient taper is taken into account, it is found that there exists an optimum illumination corresponding to maximum gain. This is shown in Fig. 6, where we plot gain, beamwidth, and the magnitude of the first side lobe. In this theoretical plot it has been assumed that the amplitude directivity pattern of the horn can be approximated by a cosine squared function.[6]

By equating our phase error at the aperture edge to the permitted criteria and expressing the lens aperture in terms of the optimum half-power beamwidth (Fig. 6), we obtain rather useful formulas expressing the number of beamwidths scanned for each lens type as a function of the beam-width and the f/D ratio of the system. These are plotted in Fig. 7 where the curves indicate the *total* scanning angle for the commonly used f/D ratio of unity. For comparison, the scanning possibilities of a parabolic cylinder, with the feed constrained to the focal arc is plotted on the same graph. The formula indicated in this case is a commonly accepted one and may be ob-

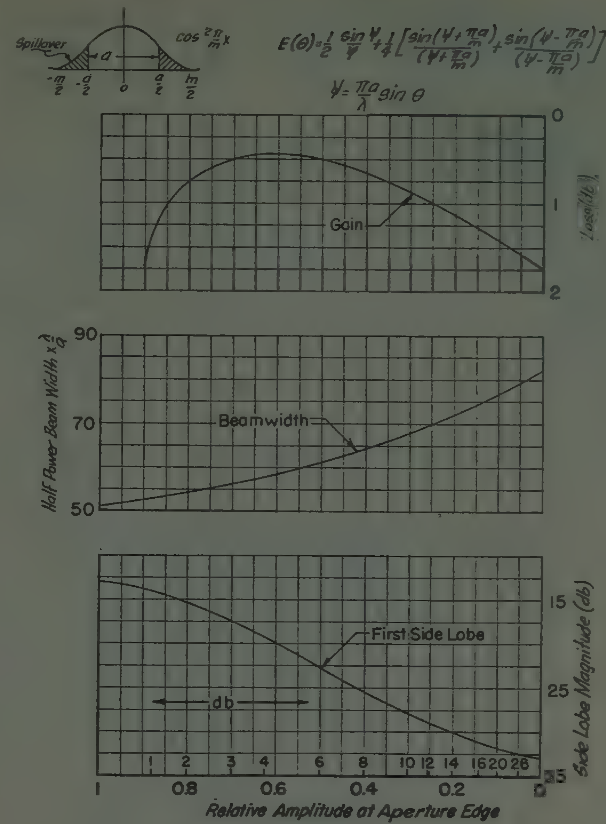


Fig. 6—Radiation characteristics of tapered rectangular aperture.

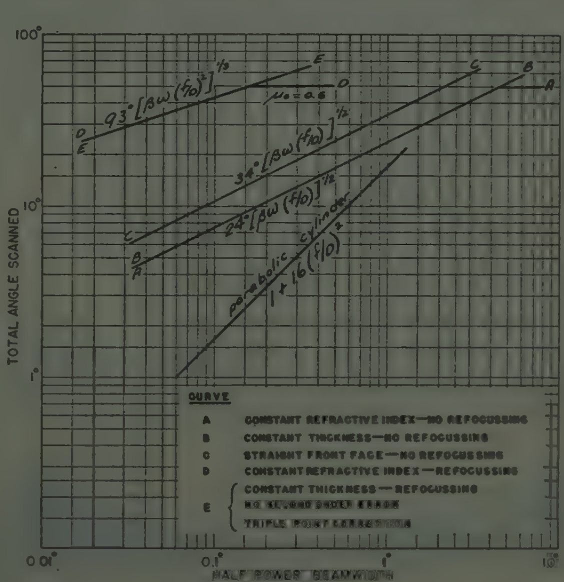


Fig. 7—Maximum scanning angle of constrained lenses.

[5] J. Ruze, "Wide Angle Metal Plate Optics," Cambridge Field Station Report No. E 5043 Air Matériel Command, March, 1949.

[6] S. Seeley, "Microwave antenna analysis," Proc. I.R.E., vol. 35, pp. 1092–1095; October, 1947.

tained by a similar phase analysis and with the same phase-error criteria as used with the lenses. The different functional form for the different types should be noted. The data presented in this figure were calculated on the assumption that optimum correction points for maximum scanning angle were used in each case.

In the case of the refocussed "straight-front-face" lens this simpler analysis indicates complete cancellation of coma. Actually, higher-order coma exists, in addition to higher-order aberrations. This lens has the best wide angle scanning possibilities; however, because of the failure of our approximate analysis, it could not be included in Fig. 7. The characteristics of this lens and others at large angles may still be determined by direct application of the exact formulas (8) and (11). This attack is illustrated in Fig. 8, where the phase error across the aperture is plotted at various feed positions for a "no-second-order" lens with correction points at ± 50 degrees. This curve indicates that as far as phase considerations are concerned it is possible to scan a total angular region of 110 degrees by the following systems:

(1) A 2-degree beam with an $f/D=1.0$
(2) A 1.2-degree beam with an $f/D=1.5$
(3) A 0.75-degree beam with an $f/D=2.0$.

A quarter-wavelength phase error was permitted at the aperture edge and no refocussing would be necessary.

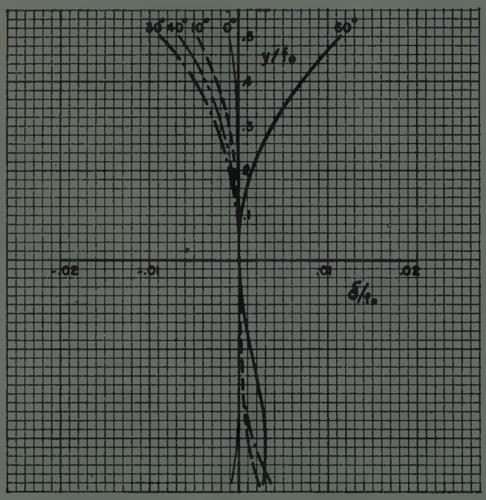


Fig. 8—Phase deviation of no-second-order lens (corrected at ± 50 degrees).

B. Shading Effect of Steps and Lens Edge

All lenses constructed were line sources suitable to illuminate a cylindrical parabola or another lens. Fig. 9 shows the construction used. The first lenses were of the constant-refractive index type with tolerance stepping in lens thickness on the outer lens contour. Although such lenses possessed predicted scanning performance, their radiation patterns illustrated a defect not expected at first. Fig. 10 shows a pattern of a 36-inch K-band (1.25 cm), stepped lens with a refractive index of 0.6. It should be noted that far side lobes of appreciable magnitude exist between 15 and 60 degrees. These side lobes are due to shading and reflection of energy at the steps. In comparison, Fig. 11 shows a pattern for a "no-second-order" lens of the same size. This lens has a variable refractive index and a smooth outside face. The pattern obtained is exceptionally clear and illustrates the importance of a smooth radiation aperture. Increasing the illumination taper resulted in a much less marked difference in the two patterns.


Fig. 9—A 72-wavelength, 1.25-cm lens.

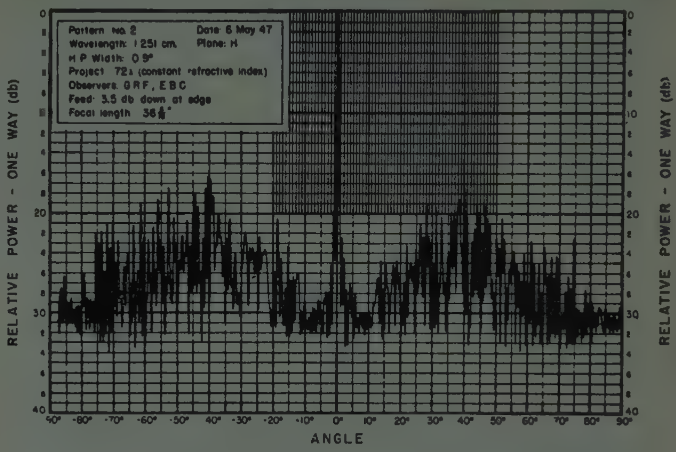


Fig. 10—Radiation pattern (stepped constant refractive index lens).

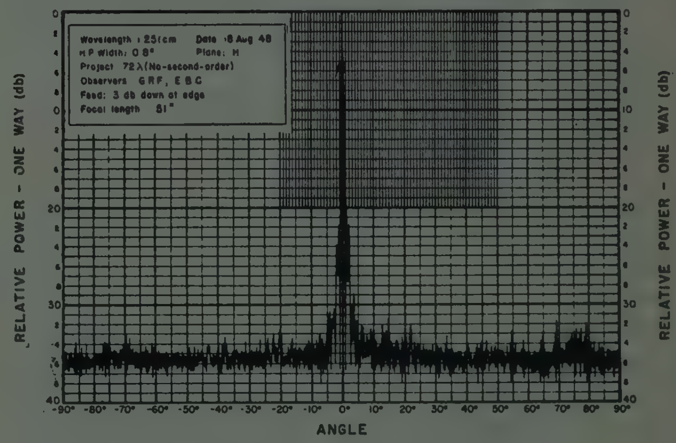


Fig. 11—Radiation pattern (no-second-order lens).

C. A Wide-Angle Double-Media Lens

The variable-refractive-index lens requires that the plate spacing vary from point to point. This may be mechanically inconvenient. It does not matter how the electrical phase change in the length "d" is accomplished. If desired, it may be achieved by combining two regions of different but constant refractive indices in the correct proportions.

A 72-wavelength, double-medium, no-second-order lens with correction points at $\pm 45°$ was investigated. The double-media action was obtained by filling a portion of the guide with polystyrene. The radiation characteristics of this lens were determined experimentally and are summarized in Fig. 12. It should be noted that acceptable patterns are obtained over a 100-degree sector. An f/D ratio of 1.5 was used.

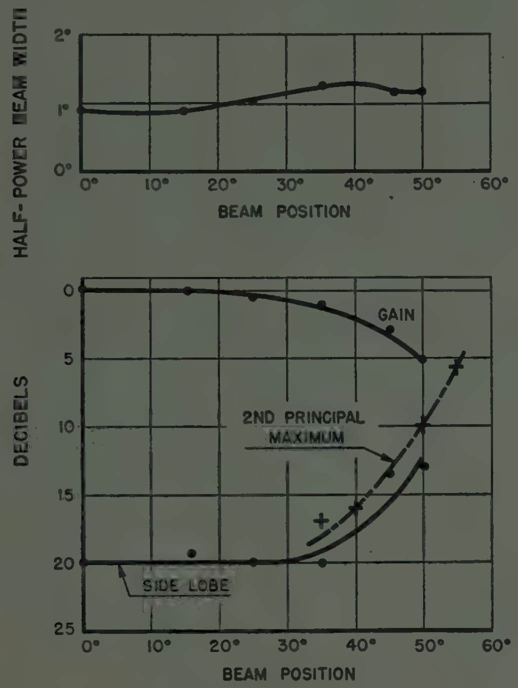


Fig. 12—Summary of data on 100-beam-width lens.

In building wide-angle lenses, care must be taken to eliminate the appearance of secondary principal maxima in the radiation pattern. These lobes appear at high angles, increase in magnitude, and rapidly move toward the center as the scanning angle is increased. They may be explained by considering the lens as a number of discrete sources, each properly phased and located at the center of each set of conducting plates. We obtain for the radiation pattern for this configuration[7]

[7] G. C. Southworth, "Certain factors affecting the gain of directive antennas," Proc. I.R.E., vol. 18, pp. 1502–1537; September, 1930.

$$G(\theta) = F(\theta) \frac{\sin N\pi A(\sin\theta - \sin\theta_0)}{N \sin \pi A(\sin\theta - \sin\theta_0)}. \tag{14}$$

where

$G(\theta)$ = field strength pattern of the array
$F(\theta)$ = field strength pattern of an individual element
(θ) = general angle from the normal
θ_0 = angle of the main beam
N = number of elements
A = spacing of elements in wavelengths.

Principal maxima of this pattern occur at

$$\sin\theta - \sin\theta_0 = \pm \frac{n}{A}, \tag{15}$$

where n is any positive integer. With the exception of the element factor all the principal maxima are of the same size. The case where n is equal to zero corresponds to the main or desired beam. Between the principal maxima there exist $N-2$ secondary maxima. These are the familiar low-level secondary lobes of a broadside antenna array.

We note that if we desire only one principal maximum (15), then the element spacing must satisfy the relationship

$$A \leqq \frac{1}{1 + \sin\theta_0}. \tag{16}$$

Equation (16) indicates that for a total scanning angle of 100 degrees it is necessary to maintain the element spacing less than 0.565 wavelength. We conclude that for wide-angle lenses we must use closely spaced elements or compact structures.

Acknowledgments

The design equations for a constant-refractive-index, double-correction-point lens were first derived in the summer of 1946 by V. A. Counter of the Antenna Laboratory of the United States Air Force Cambridge Research Laboratories, who also suggested the variable-index triple-correction-point lens. Further acknowledgment must also be made to other members of this Laboratory who contributed to this work.

High-Power Sawtooth Current Synthesis from Square Waves*

HEINZ E. KALLMANN†, SENIOR MEMBER, IRE

Summary—Steps towards a new and more efficient method of generating scanning sawtooth currents are discussed. A sawtooth wave S, with rounded peaks but with a straight slope extending over nearly 90 per cent of the period, is obtained by the following steps:

1. Synthesis of a step wave from three or four members of a series of square waves P_n whose amplitudes and periods decrease with 2^n, namely,

$$S = P_1 + 0.5P_2 + 0.25P_4 + 0.125P_8 + \cdots + 1/2^nP_{2^n};$$

and by

2. Smoothing the step wave by attenuating its higher harmonics in a simple low-pass filter network with negligible transmission at, and beyond, the lowest missing harmonic. Networks having good transient response are shown to be most suited.

The efficiency of a tube circuit for this sawtooth current synthesis approaches 50 per cent when comparing power output with power supplied to the tube anodes. A modification of this circuit, using gated square waves, has an efficiency approaching 100 per cent.

SAWTOOTH CURRENTS are now widely used for the scanning deflection of cathode-ray tube beams, particularly in television. Yet the wave forms produced by the usual sawtooth current generators are often poor, and the efficiency is very poor. An efficient generator would (1) require few components, and (2) waste little of the power supplied to it. A generator satisfying only the former condition may yet be acceptable for low power output; a high-power generator should at least satisfy the second condition. As indicated by the title, the arrangements here discussed belong to the latter group.

Any periodic wave form, according to Fourier, may be synthetized from a harmonic series of sinewaves. For the sawtooth wave S_1, this series, in a convenient form, is

$$S_1 = \sin \omega_1 + 1/2 \sin \omega_2 + 1/3 \sin \omega_3 + 1/4 \sin \omega_4 + 1/5 \sin \omega_5 + \cdots + 1/n \sin \omega_n. \tag{1}$$

In this series, there are terms for every harmonic of ω_1, and with rather slowly decreasing amplitudes. Thus, approximation of a sawtooth wave by synthesis from sinewaves is unattractive, since it can be shown that even a scanning wave straight over 85 per cent and with rounded peaks would still require at least 15 terms of the Fourier series, each to be produced with low distortion by a sinewave generator of proper amplitude and phase.

Synthesis from square waves, on the other hand, appears to be attractive, for two reasons:

1. Square waves can be produced from direct current simply and very efficiently, particularly if extreme steepness of transition is not required.

2. Only a few square waves are needed, since each square wave itself contains numerous harmonics of appreciable amplitude, as is seen from its Fourier series, conveniently written for a square wave P_1 of the frequency ω_1

$$P_1 = \sin \omega_1 + 1/3 \sin \omega_3 + 1/5 \sin \omega_5 + 1/7 \sin \omega_7 + 1/9 \sin \omega_9 + \cdots + 1/n \sin \omega_n. \tag{2}$$

Thus the square wave P_1 alone already contains all the odd terms of the sawtooth wave S_1, and with the proper amplitudes and phases.

It has been found possible to synthesize a variety of periodic wave forms from square waves to any desired degree of accuracy, particularly wave forms having one or two straight slopes per period. The following analysis will only deal with the synthesis of sawtooth waves, as the most important application of the method, and in particular of scanning waves, having rounded peaks and finite "return" time. Their synthesis takes three steps: (1) synthesis of a stepped sawtooth wave approximation from square waves; (2) smoothing of this step wave by a gradually cutting low-pass filter; (3) increasing the efficiency of this method by gating the component square waves.

Sawtooth Wave Synthesis from Square Waves

How a sawtooth wave may successively be approximated by a series of square waves is shown graphically in Fig. 1. The lowest member of the series, P_1, is of the same frequency as S_1 and of one-half its amplitude; the higher members of the series have periods and amplitudes decreasing with 2^n. Successive steps of the approximation are shown as successively finer broken lines for the first three steps, and as a solid 16-step curve for the sum of the first four square waves of the series. Further approximations, using n square waves, would approach the ideal straight sawtooth S_1 more closely, with step curves of 2^n steps. The algebraic addition corresponding to Fig. 1 is tabulated in Table I for the first 16 harmon-

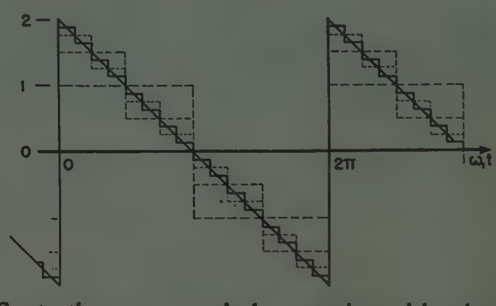


Fig. 1—Sawtooth wave successively approximated by the addition of square waves.

* Decimal classification: R355.914.432. Original manuscript received by the Institute, March 24, 1949. Presented, 1949 National IRE Convention, New York, N. Y., March 9, 1949.
† Consulting Engineer, New York, N. Y.

TABLE I
HARMONICS OF FOUR SQUARE WAVES ADDED TO APPROXIMATE HARMONICS OF A SAWTOOTH WAVE

	sin ω_1	sin ω_2	sin ω_3	sin ω_4	sin ω_5	sin ω_6	sin ω_7	sin ω_8	sin ω_9	sin ω_{10}	sin ω_{11}	sin ω_{12}	sin ω_{13}	sin ω_{14}	sin ω_{15}	sin ω_{16}
$P_1=$	1		+1/3		+1/5		+1/7		+1/9		+1/11		+1/13		+1/15	
$P_2=$		+1/2				+1/6				+1/10				+1/14		
$P_4=$			+1/4									+1/12				
$P_8=$								+1/8								
$S_1=$	1	+1/2	+1/3	+1/4	+1/5	+1/6	+1/7	+1/8	+1/9	+1/10	+1/11	+1/12	+1/13	+1/14	+1/15	+1/16

ics of the sawtooth wave S_1, and the corresponding harmonics of four square waves P_1; P_2; P_4; and P_8. The table shows that, with the amplitudes of the square waves chosen in the proportion 1:0.5:0.25:0.125, the approximation is perfect for the first 15 harmonics of the Fourier series S_1, and that the 16th is the lowest missing harmonic. Adding a square wave $1/16\ P_{16}$ would make the approximation perfect to including the 31th harmonic, and in general, for n square waves, to including the (2^n-1)th harmonic. The summation of a series of square waves to a sawtooth wave is expressed by

$$S_1 = P_1 + 1/2P_2 + 1/4P_4 + 1/8P_8 + 1/16P_{16} + \cdots + 1/2^nP_{2^n}. \quad (3)$$

This may be proved by subtracting P_1 of (2) from S_1 of (1) with the result

$$S_1 - P_1 = 1/2S_2 = 1/2P_2 + 1/4P_4 + 1/8P_8 + \cdots + 1/2^nP_{2^n}, \quad (4)$$

and repeating such subtraction for each P_n, the smaller and smaller residual sawtooth wave S_n being the difference between the step waves and the ideal sawtooth wave S_1.

Square-wave currents may conveniently be generated by switching a direct current on and off, by means of relays, or gas-filled tubes, or vacuum tubes each controlled, for instance, by a multivibrator. Since at the instant of the steep rise of the sawtooth wave all square-wave generators are to be simultaneously switched on this instant suggests itself for synchronization of all multivibrators.

The voltage drop in the switching devices will be negligible in the case of relays, a few volts in the case of gas-filled tubes, and will be relatively small even in the case of vacuum tubes, because nonlinear distortion need not be feared in the case of square waves and the tubes may be driven very hard. No energy is dissipated in the tubes while their plate current is switched off, and relatively little while it is switched on because the plate potential is then low; thus large currents may be controlled by relatively small tubes.

Fig. 2 shows a tube circuit for summation of four square-wave currents. In order to eliminate the dc component from the output even if no output transformer is used, four pairs of tubes are arranged in push-pull. The control grids of each pair are controlled in push-pull by a multivibrator (merely indicated by a generator symbol), of the frequencies P_1; P_2; P_4; and P_8, respectively. The plates of all four "push" tubes are connected together and to one branch of the push-pull output trans-

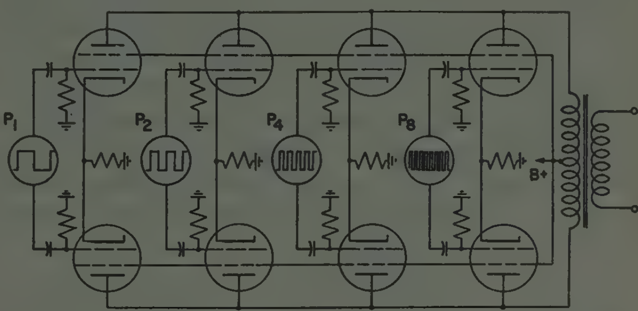


Fig. 2—Push-pull tube circuit for the addition of four square waves.

former primary; the plates of all "pull" tubes are connected to the other branch. The bias potentials are so chosen that the four pairs of plate current have amplitudes in the proportion 1:0.5:0.25:0.125. In order to simplify the circuit analysis, screen grid tubes are shown; but if triodes are substituted, the mutual interaction of their plate circuits can be cancelled by simple amplitude adjustments.

In order to complete the so obtained 16-step wave to a true sawtooth wave S_1, a sawtooth wave $1/16\ S_{16}$ might be added, of little power and with uncritical wave form, since even its second harmonic contributes only the 32nd harmonic of the wave S_1.

SMOOTHING LOW-PASS FILTER

Since the outlined method aims at the synthesis of scanning currents, particularly for television cathode-ray tubes, a true sawtooth shape with infinitely short "return" time is actually not required. Television standards permit a return time of 15 per cent of the scanning period. Such scanning wave currents can be derived from the 16-step wave currents by simple smoothing.

To find the response of suitable smoothing filters, the time response of sawtooth waves may be treated in close analogy with other time responses. In particular, it is known from the study of transient responses that ripple occurs whenever there is an abrupt change in the amplitude response of a system, and with the frequency where that abrupt change occurs. In Fig. 3 is plotted the har-

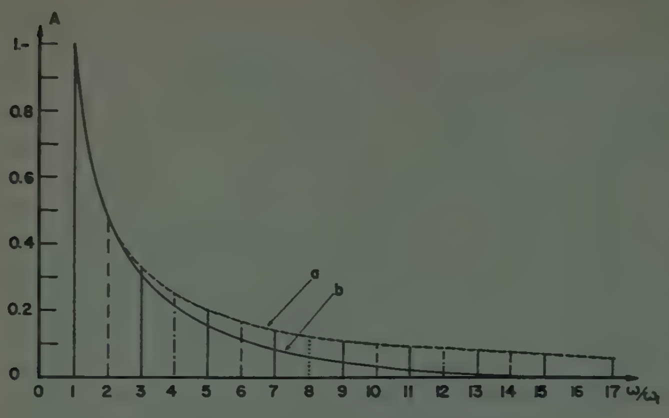


Fig. 3—Harmonic spectrum of sawtooth wave approximation before and after smoothing.

monic spectrum of the 16-step wave, according to Table I. The amplitudes of the first 15 harmonics drop steadily with frequency, with their envelope shown as the broken-line curve *a*. There is an abrupt break in this response, with the 16th harmonic the lowest and strongest missing. The 16 steps in the corresponding time-response curve may be considered as closely akin to the expected ripple of the same frequency. Indeed, addition of the 16th harmonic with the proper amplitude would fully cancel the 16-step periodicity, leaving a 32-step periodicity of weaker amplitude.

In analogy to transient response, it must be expected that a more gradual tapering off in the harmonic spectrum of the sawtooth approximation would eliminate the 16-step ripple. A thus modified amplitude response should (1) be so tapered that the amplitudes of the lowest missing harmonic, even if present, and of higher harmonics are negligible; and (2) throughout the range of appreciable amplitudes, drop so gradually that no new ripple is introduced in the time response.

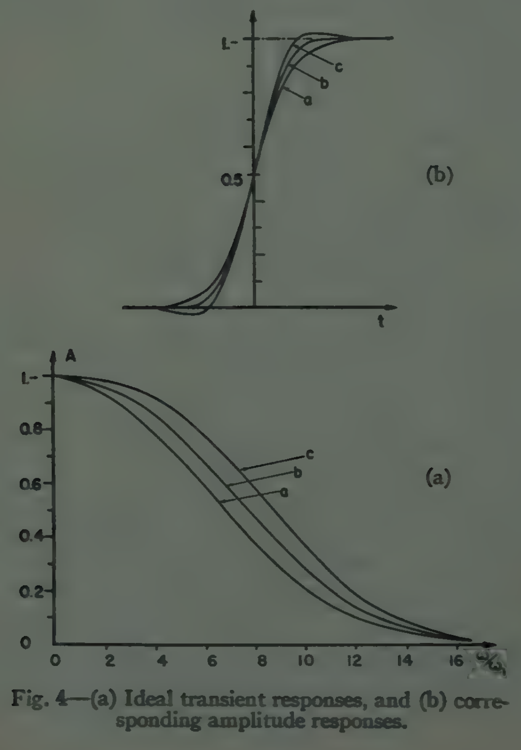


Fig. 4—(a) Ideal transient responses, and (b) corresponding amplitude responses.

The adjusted spectrum of transmitted harmonics may thus have an envelope such as the solid curve *b* in Fig. 3, obtained from curve *a* by a low-pass filter whose attenuation rises slowly with frequency. A 16-step wave corrected by such a filter will become indistinguishable from an ideal sawtooth wave that has passed through such a filter, since both were identical in all transmitted harmonics from the start, and those harmonics wherein they differed are now suppressed.

In choosing a suitable low-pass filter it is reasonable to assume—and will be confirmed below—that those filters will serve best whose time response to a unit step is smooth and free of ringing. The amplitude responses of three such filters are shown in Fig. 4(a), and the corresponding transient responses in Fig. 4(b). The transient curve *a* is the familiar ERF function, and the corresponding amplitude response curve *a* is computed from

$$A = e^{-(\omega/\omega_0)^2}. \tag{5}$$

Curves *c*, permitting a slight transient overswing, were taken from a study of transient responses,[1] and curves *b* are intermediate between *a* and *c*, with faster rise than the former but without overswing. For comparison, the frequency scale of all curves in Fig. 4(a) is adjusted for equal amplitude at ω_{16}, and the transient responses in Fig. 4(b) are scaled accordingly.

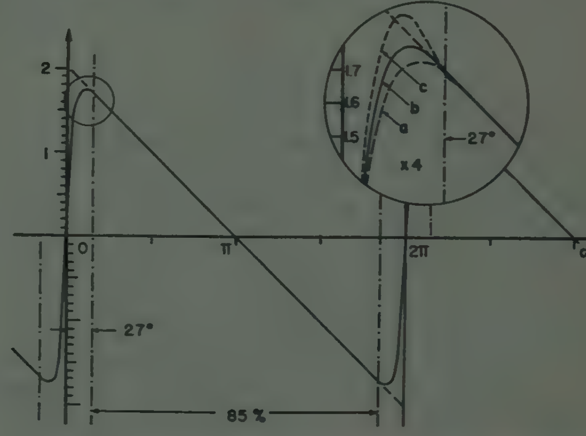


Fig. 5—Sawtooth wave approximations after smoothing.

If either an ideal sawtooth wave or its 16-step approximation are passed through such filter, a rounded sawtooth wave results as plotted in Fig. 5. With the amplitude response of each filter adjusted to about 0.02 near ω_{16}, the ripple of that frequency in the straight slope is reduced to below 0.1 per cent of the sawtooth amplitude, too small to be noticeable in the curve. In all three cases the slope is free of curvature by any standard over at least 85 per cent of the period. The steep rise is slowed and rounded, to about the same steepness in all three cases. As was the case with the three transient responses

[1] H. E. Kallmann, R. E. Spencer, and C. P. Singer, "Transient response," Proc. I.R.E., vol. 33, pp. 169–195; March, 1945. See Fig. 43, curve 3; Fig. 40, curve (2–4); and eq. (47); (54); and (56).

of Fig. 4(b), the only appreciable differences occur near the corners. To show these clearly, this part at the top of the curve, encircled in Fig. 5, is shown 4× linearly enlarged, as through a magnifier, at the right. The three peaks strikingly resemble the corners of Fig. 4(b), the broken-line curves *a* and *c* corresponding to curves *a* and *c* in Fig. 4(b), the solid curve *b* corresponding to the intermediate curve *b* in Fig. 4(b). Since none of the curves departs from the ideal over the useful 85 per cent of the period, it is evident that the amplitude response of the filter is not critical to the extent as the curves in Fig. 4(a) differ.

All the responses shown are of ideal filters and—as the symmetry of the transient responses indicates—without phase distortion. Since the very gradual cutoff required can be obtained with real filters comprising only two or three reactances, it is a simple matter to keep the phase distortion low and yet to approximate the ideal filter responses sufficiently closely. In some cases, suitably chosen stray reactances and losses in the output transformer and the deflection coils may suffice for smoothing.

The curves of Fig. 5 were computed as the sums of 15 harmonic sine waves, each attenuated according to the curves in Fig. 4(a). The same result is obtained less laboriously by arguing as follows. Since the smoothing filters are passive networks without nonlinear components, the result must be the same, whether the four square waves are added before smoothing, as done in the circuit of Fig. 2, or whether each is smoothed separately in one of four identical filters before adding. Now the shape of the separately rounded square waves can be readily predicted for all cases where the rounded portions do not "run together," i.e., where they reach their final level before another rounded portion begins. In these cases, their shape is identical with the response of the filter to a unit step, the transient response, as shown in Fig. 4(b) or to be found by familiar methods. Comparison of the time scale of Fig. 4(b) with that of Figs. 1 and 5 proves that this condition is easily met by the square waves P_1 and P_2 and barely met even by P_4, in most cases. It is not met by P_8, which indeed is reduced by the filter to a sine wave ω_8, even its next harmonic ω_{24} being suppressed.

Instead of adding the 15 harmonics, it is thus merely necessary to add three rounded square waves and one sine wave ω_8. In the most interesting range from 0° to 22.5°, the sum of the former is identical with 1.75 times the response of P_1.

Apart from simplifying computations, the argument yields the explanation why the steep rises in Fig. 5 so closely resemble the corresponding transient responses in Fig. 4(b). They are, in fact, identical in the very large part contributed by the first three square waves, and differ only in the minute part contributed by the harmonics of the fourth. It follows that, for the purpose of this study, the shape of the steep rise and of the rounded peaks of the scanning wave can be closely predicted from the transient response of the smoothing low-pass filter.

Synthesis from Gated Square Waves

When the conversion efficiency of the circuit Fig. 2 is evaluated in terms of sawtooth current versus dc supplied to the plate circuit, it is found to be below 50 per cent. Yet the losses in the smoothing filter amount only to some 3 per cent, and the voltage drop in the plate circuits of the tubes can also be held relatively low. The reason for the loss is that for much of the time some of both the "push" and of the "pull" tubes are switched on simultaneously, with currents canceling each other in the output transformer. One-half of the energy supplied is thus wasted, not dissipated in the sawtooth generator circuit itself, but in the source impedance of the power supply.

In order to avoid this loss, one may stipulate that at no time should any of the "pull" tubes be switched on simultaneously with any of the "push" tubes, thus forbidding the mutually canceling plate currents, and then try again to synthesize a sawtooth wave from square-wave components. How this may be done is shown in Fig. 6, drawn, for comparison, to the same scale as Fig.

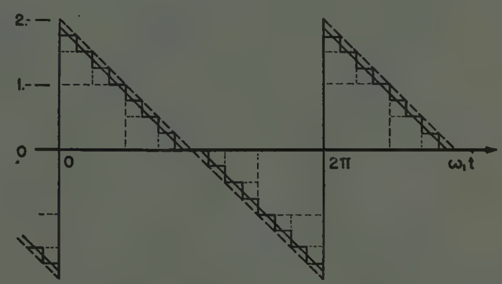


Fig. 6—Sawtooth wave approximated by addition of three gated square waves.

1. It will be noted that in Fig. 6 there are only additions, and no subtractions, of currents, and that there appears no component P_1; but there are again rectangular pulses of the square waves P_2; P_4; and P_8. These pulse sequences are shown again, sorted out, in Fig. 7; they may

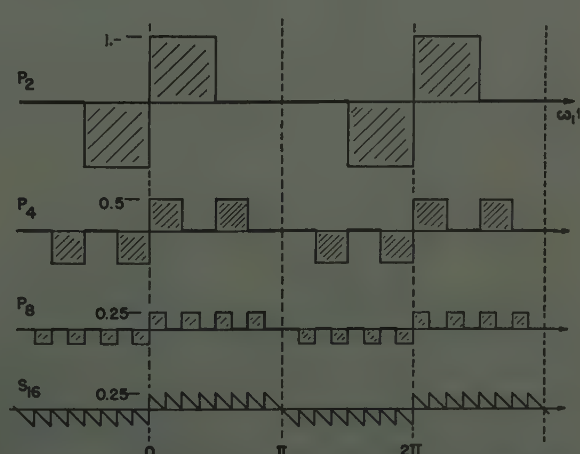


Fig. 7—Gated square wave components of Fig. 6 and supplementary gated sawtooth wave S_{16}.

be described as square waves P_2; P_4; and P_8, each gated by the square wave P_1 so that only positive pulses are permitted when P_1 is positive, from 0 to π, and only negative pulses when P_1 is negative, from π to 2π.

In the scheme of Fig. 1, the square wave P_1 supplied one-half of the sawtooth amplitude. To achieve the same amplitude, the amplitudes of the pulses P_2; P_4; and P_8 are each doubled in Fig. 6 compared with Fig. 1. But each of these currents is switched on only for one-half of its former total time, due to the gating; so that the total current through each of these tubes is not increased and, furthermore, the total current drawn from the power supply is now only one-half of that required for the same sawtooth amplitude with the original scheme of Figs. 1 and 2.

A vacuum-tube circuit suitable for pulse gating is shown in Fig. 8. It closely resembles that of Fig. 2, except that (1) the whole generator P_1 is omitted; (2) the current pulses of generators P_2; P_4; and P_8 are each

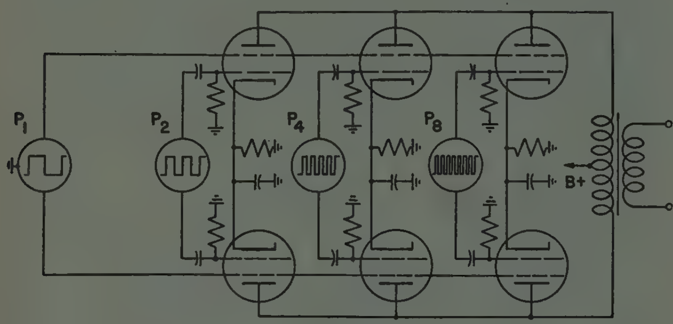


Fig. 8—Push-pull tube circuit for addition of three gated square waves.

doubled, requiring larger cathode emission; and (3) these three generators are gated by a square-wave signal P_1, so that alternatingly the "push" and the "pull" tubes are cut off. This gating may be accomplished by signals on any convenient electrode, be it the cathode, control grid, suppressor grid, etc. As shown, alternatingly positive and negative screen grid voltage is applied, with opposite phase, to the "push" and the "pull" tubes.

Again, to complete a true sawtooth wave S_1, a sawtooth wave of 6.25 per cent peak amplitude and of frequency S_{16} may be added to the square pulses, also gated as shown in the bottom line of Fig. 7; a pair of sawtooth amplifier tubes would have to be added in Fig. 8. If, however, a smooth scanning wave is desired, the previous method of smoothing by a low-pass filter is not immediately applicable. As is evident from Fig. 6, smoothing, being a process of averaging, would result in a sawtooth slope following the solid line shown instead of the desired sawtooth shown as broken line, through the corners of the 16 steps. Two faults result from this discrepancy:

1. The peak amplitude is somewhat less than in Fig. 1; this could easily be adjusted by slightly increasing all square-wave amplitudes.

2. The smoothed sawtooth approximation will have a kink in the middle of the slope, a serious fault.

The simultaneous correction of both these faults, however, happens to be simple. It is merely necessary to increase both the positive and the negative amplitude of the sawtooth approximation by a constant amplitude equal to the deficiency of the sawtooth approximation shown as solid line, compared with the desired slope shown as broken line. To this end, a shallow square wave of period P_1 is again added to the sawtooth wave approximation, modifying Fig. 6 as shown in Fig. 9. If the

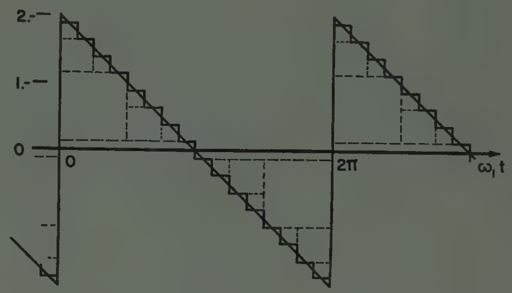


Fig. 9—Sawtooth wave approximation of Fig. 6, adapted for smoothing.

height of the "pedestal," the amplitude of the square wave P_1, is made equal to the amplitude of the first missing square wave, here P_{16}, then it can be seen that the smoothed sawtooth wave approximation will again coincide with the ideal, without kink, as in Fig. 1. A suitable modification of the circuit of Fig. 8 is shown in Fig. 10. The generator P_1 is reinserted, but now with one-eighth of the former amplitude. For similarity, its tubes are shown with the signal P_1 applied to both control grids and screen grids; this double control does not affect the result and single control would suffice.

Even in the modified circuit of Fig. 10, there are no mutually canceling currents; and its theoretical effi-

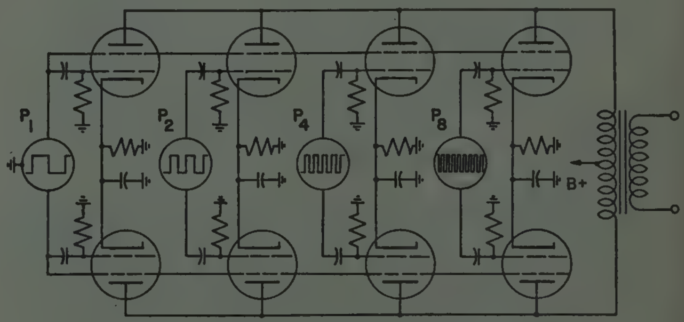


Fig. 10—Circuit as in Fig. 8, adapted for synthesis according to Fig. 9.

ciency, ignoring voltage drop in the tube plate circuits and losses in the smoothing filter, is thus 100 per cent. Yet its output may be smoothed by a low-pass filter, just as the output of the circuit of Fig. 2, and with the same result, the scanning wave shown in Fig. 5, or with real filters to a very close approximation of that wave form.

Speed of Electronic Switching Circuits*

EVERARD M. WILLIAMS†, SENIOR MEMBER, IRE, DONALD F. ALDRICH‡, ASSOCIATE, IRE, AND JAMES B. WOODFORD, JR.†, STUDENT MEMBER, IRE

Summary—**Methods of analysis of electronic switching circuits are described which lead to determination of triggering delay and switching wave forms. These methods are illustrated with particular reference to multivibrators.**

SWITCHING OR COMMUTATING circuits utilizing multivibrators, flip-flop circuits, etc., are widely used. Considerable attention has been given[1–5] to the determination of the oscillation period of these devices, but only very rough methods have been applied[1,6,7] to the study of their behavior during the switching, or transition intervals. This paper describes some more precise analyses of transition intervals with detailed application to the conventional multivibrator. The methods used, however, are adaptable to other circuits.

The analysis will be prefaced by consideration of the physical phenomena involved and some definitions involving the transition period. Fig. 1(a) is a schematic diagram of a typical multivibrator. Fig. 2(a) shows a typical output wave form; the leading edges of the traces are extremely steep and it is only with the greatly expanded horizontal time scale of Fig. 2(b) that the actual form of a leading edge appears.

During the time interval of the leading edge the output tube (T_2) grid voltage varies from cutoff to approximately zero, while the driver tube (T_1) grid voltage varies from about zero to approximately cutoff, and usually well beyond cutoff in the reverse direction. During the time interval of the trailing edge the roles are reversed. It is convenient for the purposes of analysis to divide the transition interval into three intervals:

(1) The period in which both tubes conduct and the system is driven through the loop gain. This will be termed the "driven" transition.

(2) The period during which the driver tube grid-voltage is beyond cutoff so that this tube draws no plate-current, but its anode voltage is increasing because of the charging of its output shunt capacitance through the plate load resistance. During this period

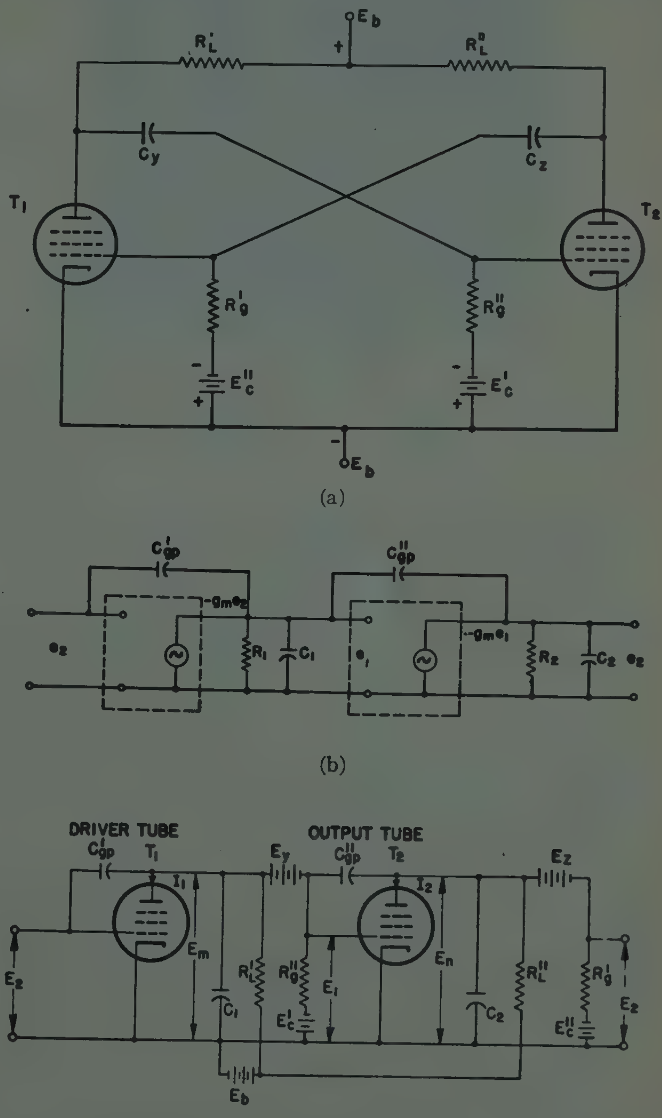


Fig. 1—(a), Schematic diagram of a typical multivibrator. (b) Equivalent circuit of this multivibrator during transition interval when the period of oscillation is long relative to transition time. Shunt capacitances (tube, stray, etc.) and resistances are combined in R_1, C_1, R_2, and C_2. The two grid-plate capacitances are effectively in parallel and their sum will be replaced by C_f. (c) Simplified circuit for step-by-step analysis for the case in which coupling capacitor voltage changes are negligible during the transition interval. The coupling capacitors are replaced by constant voltages equal to the voltage immediately before the switching interval.

* Decimal classification: R371.51×R257.2×R146.2. Original manuscript received by the Institute, May 24, 1949; revised manuscript received, October 24, 1949. Presented, 1949 IRE National Convention, New York, N. Y., March 10, 1949. This paper contains in part a section of a dissertation by James B. Woodford, Jr., as partial fulfillment of the requirements for the degree of Doctor of Science at Carnegie Institute of Technology.

† Carnegie Institute of Technology, Pittsburgh 13, Pa.
‡ Research Laboratory, Westinghouse Electric Corporation, East Pittsburgh, Pa.

[1] M. V. Kiebert and A. F. Inglis, "Multivibrator circuits," PROC. I.R.E., vol. 33, pp. 534–539; August, 1945.

[2] E. H. B. Bartelink, "A wide-band square-wave generator," *Trans. AIEE*, vol. 60, pp. 371–376; 1941.

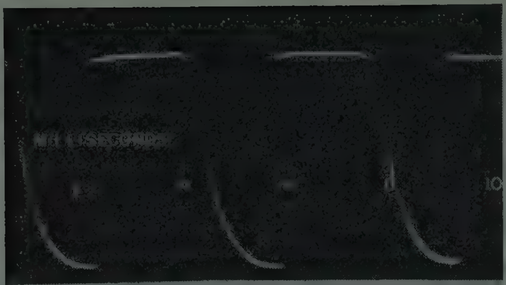
[3] R. M. Page and W. F. Curtis, "The Van der Pol four-electrode tube relaxation oscillator circuit," PROC. I.R.E., vol. 18, pp. 1921–1929; November, 1930.

[4] A. E. Abbot, "Multivibrator design by graphical methods," *Electronics*, vol. 21, pp. 118–120; June, 1948.

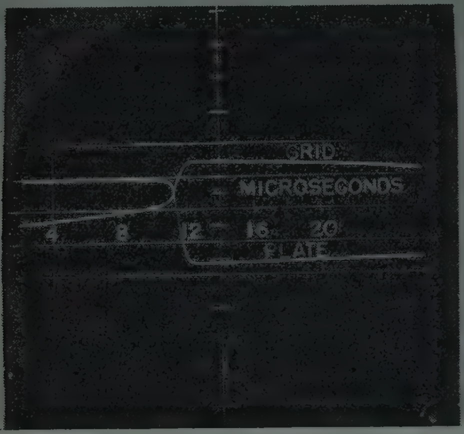
[5] R. Feinberg, "On the performance of push-pull relaxation oscillator (Multivibrator)," *Phil. Mag.*, vol. 39, pp. 268–281; April, 1948.

[6] D. G. Fink, "Radar Engineering" pp. 315–316, McGraw-Hill Book Co., New York, N. Y., 1947.

[7] David Sayre, "Generation of Fast Waveforms," Chapter V of "Waveforms," edited by B. Chance, McGraw-Hill Book Co. New York, N. Y., 1949.



(a)



(b)

Fig. 2—(a) Typical output voltage of the multivibrator of Fig. 1 (a). (b) View of output tube grid and plate-to-cathode voltages during a short interval about the leading edge of the wave form in (a) above.

the output tube grid voltage is increasing and its output capacitance is discharging. This period will be termed the "intermediate" portion of the interval.

(3) The period during which the output tube grid voltage is constant at a value slightly greater than zero, hereafter called a "quasi-stable" value, while the output circuit shunt capacitance continues to discharge through the output tube anode circuit. The interval, which will be designated the "terminal" portion, ends when the output tube plate current drops to the value determined by the output load resistance.

The driven interval is necessarily always present, but there may not be a recognizable intermediate or terminal interval. Since the trailing transition is described by interchanging the roles of the driver and output tubes, these same intervals may also be identified in a trailing transition although the actual output can depend, at most, upon the first two. The term "transition interval" will be understood to include all intermediate periods.

The influence of various circuit parameters during the "driven" interval can be approximately determined by a simple analytical method. The approach used is to represent the tubes by equivalent circuits, in which "mean" values of tube transconductances and plate resistances are employed to allow for their wide variation over the operating range. The equivalent circuit of Fig. 1(b) will be used for illustration. In this circuit it is assumed that the coupling network time constants (coupling capacitors and associated plate load resistor, grid resistors, grid input resistance) are long relative to the shunt network time constants, which depend on output and input tube capacitances, stray capacitances to ground, grid and plate resistances, etc. This assumption is valid unless the oscillation periods are very short; in such cases a more elaborate analysis is necessary. With this assumption the effect of the coupling capacitance may be neglected during the transition period. The illustrative work of this paper will, in addition, be simplified by assuming equal tube transconductances.

The differential equations for e_1 and e_2 are identical in form for the driven interval. This form is

$$[C_1C_2 + C_f(C_1 + C_2)]\frac{d^2e}{dt^2} + \left[\frac{(C_1 + C_f)R_1 + (C_2 + C_f)R_2}{R_1R_2} + 2C_fg_m\right]\frac{de}{dt} + \left[\frac{1}{R_1R_2} - g_m\right]^2 e = 0.$$

The solution is

$$e_1 = A\epsilon^{(\beta-\alpha)t} + B\epsilon^{-(\beta+\alpha)t}$$
$$e_2 = D\epsilon^{(\beta-\alpha)t} + F\epsilon^{-(\beta+\alpha)t},$$

in which

$$\alpha = \frac{R_1C_1 + R_2C_2 + C_f(R_1 + R_2) + 2C_fg_mR_1R_2}{2R_1R_2(C_1C_2 + C_f[C_1 + C_2])} \tag{1}$$

$$\beta = \left[\alpha^2 + \left(g_m^2 - \frac{1}{R_1R_2}\right)\left(\frac{1}{C_1C_2 + C_f(C_1 + C_2)}\right)\right]. \tag{2}$$

The transition interval is initiated by a triggering pulse, or by the voltage decay at the output tube grid due to discharge of the coupling capacitance. We may write either condition in terms of the rate of change of grid voltage de_1/dt at the instant of triggering, as

$$t = 0 \qquad e_1 = 0 \qquad \frac{de_1}{dt} = \left(\frac{de_1}{dt}\right)_{t=0},$$

which gives for e_1

$$e_1 = \frac{1}{2\beta}\left(\frac{de_1}{dt}\right)_{t=0}\epsilon^{(\beta-\alpha)t} - \frac{1}{2\beta}\left(\frac{de_1}{dt}\right)_{t=0}\epsilon^{-(\beta+\alpha)t}.$$

For all normally encountered values of $(de_1/dt)_{t=0}$ the value of $1/2\beta(de_1/dt)_{t=0}$ is of the order of magnitude of unity, or less, so that for appreciable voltages in the driven transition

$$e_1 = \frac{1}{2\beta}\left(\frac{de_1}{dt}\right)_{t=0}\epsilon^{(\beta-\alpha)t} \tag{3}$$

and the significant factors in e_2 are contained in

$$e_2 = \frac{g_m}{2C_2(\beta - \alpha)}\left[\frac{de_1}{dt}\right]_{t=0}\epsilon^{(\beta-\alpha)t}. \tag{4}$$

The voltages e_2 and e_1 may be regarded as the steep parts of the leading and trailing transitions, respectively. The solutions yield an indefinitely rising voltage, but are no longer applicable after the driver tube grid voltage drops below cutoff or the output tube grid voltage increases above zero to the quasi-stable value; either event terminates the driven interval. The time duration T_d of the driven interval is given by the solution of (3) or (4) for the time required for the voltage to rise to the value which terminates the interval. If we designate the critical grid voltages E_a (cutoff of driver tube) and E_b (quasi-stable value of output tube), respectively, the time T_d is given by the shorter of the two times:

$$T_d' = \frac{1}{\beta - \alpha} \log_e \frac{2\beta E_a}{\left(\frac{de_1}{dt}\right)_{t=0}}, \tag{5}$$

or

$$T_d' = \frac{1}{\beta - \alpha} \log_e \frac{2\beta E_b C_2(\beta - \alpha)}{g_m \left(\frac{de_1}{dt}\right)_{t=0}}. \tag{6}$$

Owing to the origin of these expressions, we can assume that the time values computed by their means are approximate, but that they should reflect with some accuracy the effect of changing circuit parameters. They have, in fact, been used by the authors for this purpose with considerable success.

The utility of the approximate approach will be illustrated in detail for the effect of triggering voltage.

Equations (3) and (4) show that the time interval for the voltage to change from any particular voltage E_d to another E_f is given by

$$\Delta t = \frac{1}{\beta - \alpha} (\log_e E_f - \log_e E_d), \tag{7}$$

while the delay time T_d between a triggering pulse and a transition to a particular voltage E_k is given by

$$T_d = \frac{1}{\beta - \alpha}\left[\log_e E_k + \log_e 2\beta - \log_e \left(\frac{de_1}{dt}\right)_{t=0}\right]. \tag{8}$$

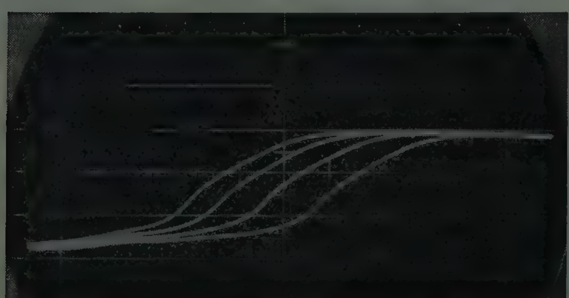


Fig. 3—Superposed traces of transition voltage for four values of triggering voltage in the ratios (left to right) 5, 4, 3, 2. The triggering pulse also appears, slightly to the left of the first transition.

Equation (7) indicates that the shape of the voltage in the transition interval should be independent of the triggering voltage, and (8) that the delay should decrease linearly with the logarithm of the rate of change of triggering voltage. Fig. 3 shows the superposed

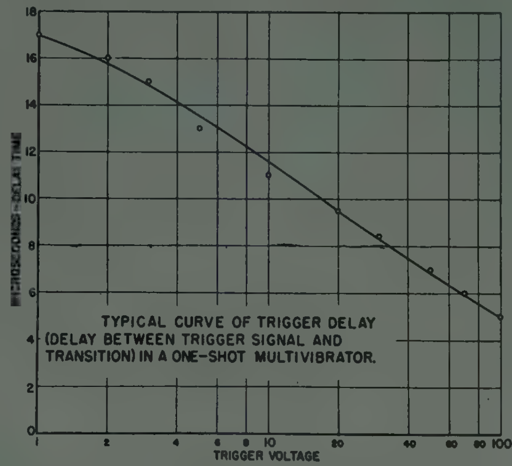


Fig. 4—Trigger delay in a one-shot arrangement of a conventional multivibrator as a function of triggering voltage, showing close approximation to a straight-line relationship. The circuit for which these data were taken was operated in such a way that the shape of the triggering voltage was independent of amplitude and $(de_1/dt)_{t=0}$ was thus proportional to the amplitude.

traces of transitions in which a wide range of triggering voltages were used, demonstrating that there is no appreciable effect on the shape of the driven portion of the transition (or indeed, upon the whole transition) for different values of $(de_1/dt)_{t=0}$. Fig. 4 demonstrates that the delay in a triggered multivibrator is quite closely dependent on the logarithm of the trigger voltage (or on the logarithm of the rate of change of trigger voltage since this is proportional to trigger voltage). It is also apparent that if the delay is known for two values of triggering voltage, it may be readily calculated within wide limits for other triggering voltages.

The voltage variation during a driven transition can be calculated with considerable accuracy by a step-by-step method, applicable only when specific values of tube characteristics and circuit parameters are known. The step-by-step attack offers the only means by which sufficiently accurate boundary values can be determined to provide initial data for calculating performance during the intermediate and terminal intervals. This technique is described in the Appendix.

Conclusions

Several general conclusions, which have been checked experimentally, have been drawn from application of the methods of this study. Among these are:

(a) There seems to be little correlation between transition time and the time constants of the load resistances (in parallel with dynamic plate resistance) in combination with shunt capacity, although this has been suggested as a major factor by Kiebert and Inglis and others.[1,6] The shunt capacity is indeed of major

importance, but the load resistance of the driver stage appears important primarily because low values of this load resistance result in higher values of quasi-stable output tube grid voltage (and shorter terminal intervals) during the terminal interval, and the load resistances of both driver and output stages seem generally significant only as second-order effects in determining the transconductance range in which the tubes operate during the driven interval. Very large values of load resistance will reduce the mean transconductance and result in longer driven intervals and greater triggering delays. The contradiction between this general conclusion as to the role of load resistance and the conclusions of earlier investigators would appear to arise from their assumption that the vacuum tubes switch from cutoff to conduction, or vice-versa, instantaneously, and that finite transitions in output voltage arise from the charging time constants of shunt circuits associated with the vacuum tube.

(b) When the periods of multivibrator oscillations are of the same order of magnitude as the transition time, the transition times are increased, since the changes of voltage across coupling capacitors, appreciable during the transition interval in this case, have a degenerative effect.

(c) The parameter which appears of greatest significance in determining switching speed is the figure-of-merit gm/C for the tubes used.

The methods of this paper have also been applied to other types of switching circuit, in particular, the Eccles-Jordan, and cathode-coupled multivibrators. These circuits cannot be simplified for analytical purposes as readily as the conventional multivibrator, and the resulting equations are not as generally useful. The step-by-step methods, however, can be applied to these circuits without much additional complication.

Appendix

Step-by-step methods are applicable to all portions of a transition interval. The procedure followed is illustrated in detail for the driven interval.

1. Driven Interval

The driven interval is initiated when the output tube grid voltage rises above cutoff. The assumption (used earlier in the approximate analytical analysis) that the coupling network time constant is long relative to the transition interval is usually valid, and will be applied here for simplicity, although not necessary for the application of graphical methods. The circuit then reduces to that of Fig. 1(c), similar to the equivalent circuit of Fig. 1(b), except that fixed voltages as well as varying components are included.

Initially, at a time t_0 immediately before triggering, the grid voltage E_1 is either at cutoff, if the bias E_c' is less than cutoff, or at the voltage E_c' if E_c' is greater than cutoff. If sufficient time has elapsed since the previous transition

$$E_n = E_b$$

$$E_2 = E_c'',$$

and E_m and I_1 are at values which may be determined graphically from the characteristics of the drive tube T_1 by using the load line for a load R_L'. Furthermore, the coupling capacitors are charged to voltages

$$E_2 = E_B + E_c''$$

$$E_y = E_m + (E_1)_{\text{initial}},$$

and these voltages will remain constant throughout the switching interval. The relative importance of grid-plate capacitances may be determined by solving for α and β from (1) and (2) with and without grid-plate capacitance; the process to be described is greatly simplified if these capacitances can be omitted from the equivalent circuits.

From a knowledge of the triggering voltage, external or internal, a value of $(E_1)_{t_0+\Delta t}$ following a small time interval Δt after reference time, t_0, can be determined. A precise knowledge of triggering voltage is of concern only when the triggering delay must be calculated accurately, since, unlike delay, transition voltage form is independent of the magnitude of the triggering voltage (except for the case in which the triggering voltage continues during the transition interval itself and is of such magnitude and shape as to seriously distort the transition). From the value of $(E_1)_{t_0+\Delta t}$ a value of anode current $(I_2)_{t_0+\Delta t}$ may be determined from the characteristics of tube T_2 using the initial value of $E_n = E_b$. The output voltage E_2 of the linear coupling network between T_2 and T_1 is then calculated for an input step impulse of current of the magnitude $(I_2)_{t_0+\Delta t} - (I_2)_{t=0}$. A value of $(I_2)_{t_0+3\Delta t} \cdots$, etc., must also be calculated for use as a boundary condition in a subsequent step. This output voltage is calculated for time $t_0+2\Delta t$ and

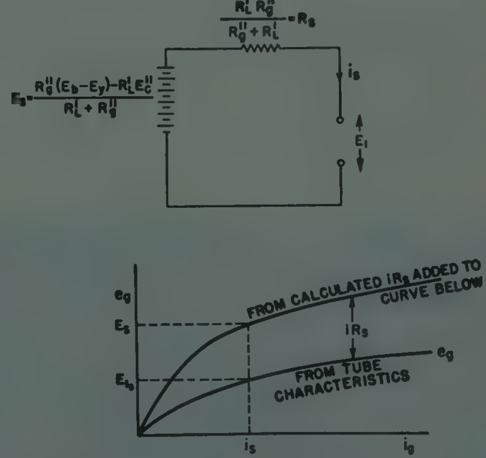


Fig. 5—Equivalent circuit used in determining quasi-stable value of E_{1_s}, or maximum positive value of grid voltage that E_1 can assume. Since this is usually a low value, the effect of E_n on the grid current curve can usually be neglected. The effect of tube T_1 is also neglected, since its grid voltage is likely to be near, at, or below cutoff when E_1 assumes its final value.

applied to tube T_1 to determine a current $(I_1)_{t_0+2\Delta t}$, which is applied as a step-current function to the coupling network between T_1 and T_2 to determine a new value $(E_1)_{t_0+3\Delta t}$. This step-by-step process is continued until the end of the driven interval, which occurs either when E_2 drops below the cutoff value for T_1 (a readily recognizable condition) or when E_1 reaches a quasi-stable positive value E_{1_s} limited by grid-current in T_2. This quasi-stable voltage limit may usually be calculated with sufficient accuracy from the equivalent circuit of Fig. 5. If E_{1_s} is reached before the driver tube plate current is cut off, there is no intermediate interval; this is a condition likely to occur in short grid-base, high transconductance, low-capacitance pentodes. If, on the other hand, the intermediate interval exists, but E_1 never reaches E_{1_s}, there is no terminal interval; this latter condition is common in long grid-base triodes.

2. Intermediate and Terminal Intervals

Although the calculations for E_1 and E_2 might be carried out formally by the same method as in the driven interval, the behavior of E_1 is quite independent of E_2 during these intervals and more direct methods may be devised for its solution. For instance, if grid-plate capacitances are neglected, the voltage E_1 may be determined analytically as a simple exponential function to the point at which the output tube grid-voltage becomes positive.

Reciprocity Between Generalized Mutual Impedances for Closed or Open Circuits*

A. G. CLAVIER†, FELLOW, IRE

Summary—**Let two wires *L1* and *L2* be considered and electromotive forces E_{P1} and E_{P2} be applied at two small gaps *P1* and *P2* of the wires, which can either be part of a loop or an open circuit (antennas). The two circuits are assumed to be placed in an isotropic medium. A general expression is obtained between E_{P1} and E_{P2} and the corresponding currents I_{P1} and I_{P2} at the gaps. In case the electromotive forces are sinusoidal and of the same frequency, the above expression leads to generalized self- and mutual impedances. The mutual impedances are shown to be reciprocal, the demonstration being derived from the reciprocity theorem in the form given to it by Ballantine. The formulas are applied to different cases, such as quasi-stationary current distribution, closely or loosely coupled antennas, and wave projectors. Attention is drawn to the conditions assumed in the demonstration for its validity.**

I. Definition of Generalized Mutual Impedances

LET TWO metallic systems *D1* and *D2* be considered. It will be assumed that the vector current density $\mathbf{j}$ is related linearly to the total electric field $\mathbf{e}_{\text{total}}$ at all points of the systems considered. This is expressed by the following equation

$$\mathbf{j} = \gamma \mathbf{e}_{\text{total}},$$

where γ is the local conductivity.

For the first metallic system *D1*, the total electric field may be considered as consisting of three parts:

A. An impressed electric field $\mathbf{e}_{a1}$, which is due to the action of a source of electric energy acting on *D1*.

B. A self-induced electric field $\mathbf{e}_1$, due to the action of electric charges set in motion in *D1*.

C. An electric field $\mathbf{e}_{21}$, due to the action of the moving charges on *D2*. The same applies to *D2* giving the following two equations

$$\mathbf{e}_{a1} = \frac{\mathbf{j}_1}{\gamma_1} - \mathbf{e}_1 - \mathbf{e}_{21} \tag{1}$$

$$\mathbf{e}_{a2} = \frac{\mathbf{j}_2}{\gamma_2} - \mathbf{e}_2 - \mathbf{e}_{12}. \tag{2}$$

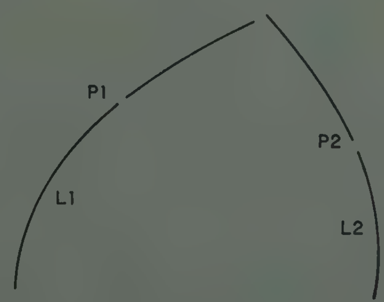


Fig. 1

By scalar multiplication of (1) by $\mathbf{j}_1$ and (2) by $\mathbf{j}_2$, two corresponding energy equations can be written.

$$\int_{D1} \mathbf{e}_{a1}\mathbf{j}_1 dv = \int_{D1} \frac{\mathbf{j}_1^2}{\gamma_1} dv - \int_{D1} \mathbf{e}_1\mathbf{j}_1 dv - \int_{D1} \mathbf{e}_{21}\mathbf{j}_1 dv \tag{3}$$

$$\int_{D2} \mathbf{e}_{a2}\mathbf{j}_2 dv = \int_{D2} \frac{\mathbf{j}_2^2}{\gamma_2} dv - \int_{D2} \mathbf{e}_2\mathbf{j}_2 dv - \int_{D2} \mathbf{e}_{12}\mathbf{j}_2 dv. \tag{4}$$

* Decimal classification: R145. Original manuscript received by the Institute, January 5, 1949; revised manuscript received, July 22, 1949. Presented, joint meeting, American Section URSI, and Washington Section, IRE, Washington, D. C., May 4, 1948.

† Federal Telecommunication Laboratories, Inc., Nutley, N. J.

Let the problem be restricted to the case of two wires with electromotive forces ξ_1 and ξ_2 applied at two infinitesimal gaps $P1$ and $P2$ (Fig. 1).

Then

$$\int_{D1} e_{a1} j_1 dv = \int_{L1} i_1 e_{a1} dl = i_{P1}\mathcal{E}_{P1}$$

$$\int_{D1} \frac{j_1^2}{\gamma_1} dv = \int_{L1} r_1 i_1^2 dl,$$

where r_1 is the linear resistance of wire $L1$ and i_1 is the current intensity. Similar transformations being applied to the terms of (3) and (4) finally result in

$$\mathcal{E}_{P1} = i_{P1}\left[\int_{L1} r_1 \frac{i_1^2}{i_{P1}^2} dl - \int_{L1} \frac{i_1}{i_{P1}} \frac{e_1 dl}{i_{P1}}\right] + i_{P2}\left[-\int_{L1} \frac{i_1}{i_{P1}} \frac{e_{21} dl}{i_{P2}}\right] \tag{5}$$

$$\mathcal{E}_{P2} = i_{P1}\left[-\int_{L2} \frac{i_2}{i_{P2}} \frac{e_{12} dl}{i_{P1}}\right] + i_{P2}\left[\int_{L2} r_2 \frac{i_2^2}{i_{P2}^2} dl - \int_{L2} \frac{i_2}{i_{P2}} \frac{e_2 dl}{i_{P2}}\right]. \tag{6}$$

Let the problem be further restricted to the case of functions varying harmonically with time so that

$$i_{P1} = \text{Real } I_{P1} \exp j\omega t$$

$$\mathcal{E}_{P1} = \text{Real } \mathcal{E}_{P1} \exp j\omega t, \text{ etc.}$$

In this case, (5) and (6) become

$$\mathcal{E}_{P1} = I_{P1}\left[-\int_{L1} r_1 \frac{I_1}{I_{P1}} \frac{I_1^*}{I_{P1}^*} dl - \int_{L1} \frac{I_1^*}{I_{P1}^*} \frac{E_1 dl}{I_{P1}}\right] + I_{P2}\left[-\int_{L1} \frac{I_1^*}{I_{P1}^*} \frac{E_{21} dl}{I_{P2}}\right] \tag{7}$$

$$\mathcal{E}_{P2} = I_{P1}\left[-\int_{L2} \frac{I_2^*}{I_{P2}^*} \frac{E_{12} dl}{I_{P1}}\right] + I_{P2}\left[\int_{L2} r_2 \frac{I_2 I_2^*}{I_{P2} I_{P2}^*} dl - \int_{L2} \frac{I_2^*}{I_{P2}^*} \frac{E_2 dl}{I_{P2}}\right]. \tag{8}$$

Equations (7) and (8) define self- and mutual generalized impedances.

$$\mathcal{E}_{P1} = Z_{11} I_{P1} + Z_{12} I_{P2} \tag{9}$$

$$\mathcal{E}_{P2} = Z_{21} I_{P1} + Z_{22} I_{P2}. \tag{10}$$

For any physical circuits, (9) and (10) are consistent, so that

$$Z_{11}Z_{22} - Z_{12}Z_{21} \neq 0.$$

There is reciprocity between mutual impedances when

$$Z_{12} = Z_{21}.$$

The present paper is concerned with the demonstration of this property and the formulation of the conditions for which it is valid.

II. Analogy with the Electrostatic Case of Mutual Coefficients of Capacitance

It is of interest at this stage to recall the demonstration of the reciprocal property of electrostatic coefficients of capacitance.

Let a closed electrostatic system be considered (Fig. 2) consisting of two metallic bodies $D1$ and $D2$ inside a metallic screen $D0$.

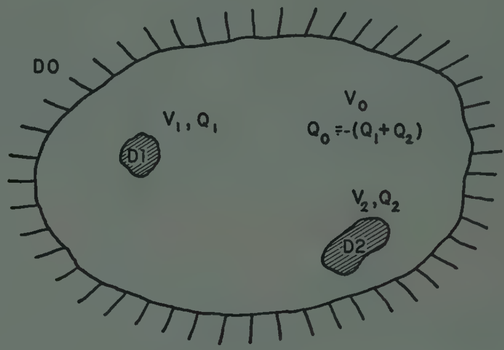


Fig. 2

The coefficients of capacitance are defined in terms of the charges and electrostatic potentials by the following equations:

$$Q_1 = C_{11}(V_1 - V_0) + C_{21}(V_2 - V_0)$$

$$Q_2 = C_{12}(V_1 - V_0) + C_{22}(V_2 - V_0).$$

To demonstrate that $C_{21} = C_{12}$, a preliminary lemma is demonstrated (lemma from Gauss) utilizing Green's theorem. Let a closed system be considered with n metallic bodies inside a metallic screen. Let two possible equilibrium states be denoted by $Q_0, Q_1, Q_2 \cdots Q_n$; $V_0, V_1 \cdots V_n$, and $Q_0', Q_1' \cdots Q_n'$; $V_0', V_1' \cdots V_n'$. It can be shown that

$$\sum [Q_n(V_n' - V_0') - Q_n'(V_n - V_0)]$$

$$= \int_S (V'E_N - VE_N) ds = \int_v \text{div } (V'E - VE') dv = 0,$$

where the $\boldsymbol{E}$ are the electrostatic field intensities in the two states, and E_N their components normal to the metallic surfaces. The notation div is for the usual operation of divergence

$$\left(\text{symbolically div} = \boldsymbol{i}\frac{\partial}{\partial x} + \boldsymbol{j}\frac{\partial}{\partial y} + \boldsymbol{k}\frac{\partial}{\partial z}\right).$$

Applying the lemma for the particular case of two bodies inside the screen and $V_1 = V_0$; $V_2' = V_0'$ results in the following equation:

$$Q_1(V_1' - V_0') = Q_2'(V_2 - V_0)$$

from which it is immediately deduced that

$$C_{21} = C_{12}.$$

III. Electromagnetic Reciprocity Theorem

Just as the demonstration of the reciprocal property of coefficients of capacitance rests with the preliminary demonstration of the lemma from Gauss, so the demonstration of the reciprocity of generalized mutual impedances rests with the demonstration of the "reciprocity theorem."

Let an electromagnetic system be considered, consisting of a certain number of metallic bodies placed in a dielectric medium characterized by constant dielectric and magnetic coefficients ϵ and μ. Suppose a certain volume V enclosed in a surface $\sum$ is considered. Consider two possible electromagnetic states, characterized by their distribution of impressed electric intensities ($\boldsymbol{E}_a$, $\boldsymbol{E}_a'$) and corresponding current densities ($\boldsymbol{J}$, $\boldsymbol{J}'$) for the particular case of sinusoidal time variations of angular velocity ω. The reciprocity theorem is then expressed by the following equation:[1]

$$\int_v [\boldsymbol{E}_a\boldsymbol{J}' - \boldsymbol{E}_a'\boldsymbol{J}]dv = \text{flux}_\Sigma \frac{1}{4\pi}[\boldsymbol{E}\times\boldsymbol{H}' - \boldsymbol{E}'\times\boldsymbol{H}],$$

where $\boldsymbol{H}$ and $\boldsymbol{H}'$ are the magnetic field intensities in the two states considered and $\boldsymbol{E}\times\boldsymbol{H}'$ denotes a vector product.

As $\boldsymbol{J}=\gamma(\boldsymbol{E}_a+\boldsymbol{E})$, where $\boldsymbol{E}$ is the electric field superposed to the impressed field by the action of the moving charges in the system, the left-hand term of the above equation reduces to

$$\int_v \gamma[\boldsymbol{E}_a\boldsymbol{E}' - \boldsymbol{E}_a'\boldsymbol{E}]dv.$$

Furthermore, the various quantities involved are related by Maxwell's equations

$$\text{curl}\,\boldsymbol{H} = 4\pi\boldsymbol{J} + j\omega\epsilon\boldsymbol{E}$$

so that the volume integral is found to be equal to

$$\frac{1}{4\pi}\int_v \text{div}\,[\boldsymbol{E}\times\boldsymbol{H}' - \boldsymbol{E}'\times\boldsymbol{H}]dv$$

equivalent, therefore, to

$$\text{flux}_\Sigma \frac{1}{4\pi}\{\boldsymbol{E}\times\boldsymbol{H}' - \boldsymbol{E}'\times\boldsymbol{H}\},$$

which demonstrates the reciprocity theorem in the form given to it by Ballantine.[1]

It can now be observed that extending the volume V to the whole space, the quantity

$$\text{flux}_\Sigma \frac{1}{4\pi}\{\boldsymbol{E}\times\boldsymbol{H}' - \boldsymbol{E}'\times\boldsymbol{H}\}$$

tends toward[2] 0.

This is so because far from the metallic bodies the fields reduce to radiated fields. At a very great distance, all points on an enclosing surface may be considered as located on a sphere with a fictitious equivalent electromagnetic source at its center. The electric and magnetic radiated fields tend to become perpendicular to the radius and are such as to produce a localized equipartition of electric and magnetic energy ($E=(\mu/\epsilon)^{1/2}H$). The theorem of reciprocity can thus be written for any physical system.[3]

$$\int_\infty \{\boldsymbol{E}_a\boldsymbol{J}' - \boldsymbol{E}_a'\boldsymbol{J}\}dv = 0.$$

Reducing this to the special case of wire circuits (Fig. 1)

$$\int_{L1} i'\boldsymbol{E}_a dl = \int_{L2} i\boldsymbol{E}_a' dl$$

or

$$\int_{L1} i d\mathcal{E}_a = \int_{L2} i d\mathcal{E}_a'.$$

Let the theorem be first applied to two closed circuits (Fig. 3). The linear dimensions of the circuits are

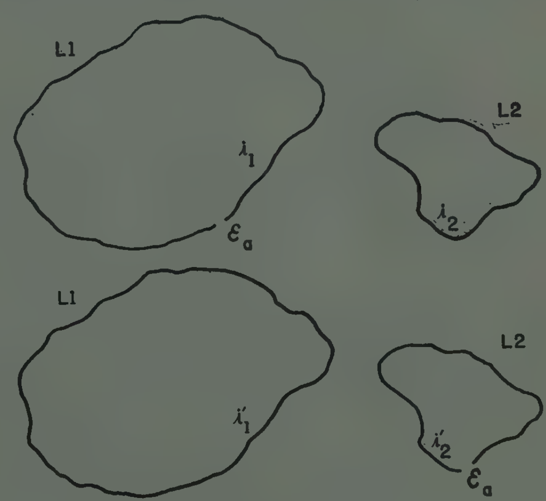


Fig. 3

assumed to be sufficiently small with respect to the wavelength, corresponding in the dielectric medium (ϵ, μ) to the angular velocity ω, so that the current intensities may be considered as constant along the wires.

In a first state, an electromotive force ϵ_a is applied to $L1$ and the currents observed in $L1$ and $L2$ are i_1 and i_2, respectively. In a second state, the same electromotive force is applied to $L2$ and the currents are i_1'

[1] Stuart Ballantine, "Reciprocity in electromagnetic, mechanical, acoustical, and interconnected systems," Proc. I.R.E., vol. 17, pp. 929–951; June, 1929.

[2] H. A. Lorentz, "Communication," *Amsterdamer Akademie van Wetenschappen*, vol. 4, p. 176; 1895–1896.

[3] J. R. Carson, "Reciprocal theorems in radio communication," *Bell Sys. Tech. Jour.*, vol. 3, pp. 393–399; July, 1924. Also Proc. I.R.E., vol. 17, pp. 952–956; June, 1929.

and i_2'. The reciprocity theorem gives

$$\mathcal{E}_a i_1' = \mathcal{E}_a i_2$$
$$i_1' = i_2.$$

This is the form given by Lord Rayleigh to the reciprocity theorem. It can be generalized to transient phenomena as the effect of any electromotive force suddenly applied to one of the circuits will be equivalent to the summation of the differential components of a Fourier integral, and the theorem will apply for each differential component individually. It can also be easily extended to all forms of passive electrical networks. An essential condition for the validity of the theorem in practical cases is, however, that should the electromotive force be produced by a generator and the current measured by an ammeter, both generator and ammeter should be such as to present a negligible internal impedance with respect to the impedances of the external circuits.[4] Thus the positions of an "impedanceless" generator and an "impedanceless" ammeter in a passive circuit may be interchanged without affecting the current through the ammeter, either in magnitude or in phase, relative to the generator voltage.[5]

The passive networks need not be of the closed type provided the positions of the electromotive force and measured current are accurately specified (Fig. 4).

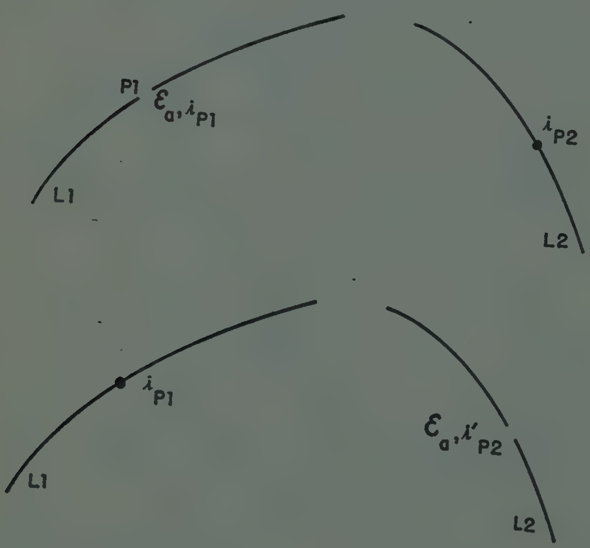


Fig. 4

If two open-wire circuits are considered, and the two states shown in Fig. 4 realized, then $i_{P1}' = i_{P2}$.

The reciprocal property of generalized mutual impedances is just another way of expressing the same thing. In the first state,

$$\mathcal{E}_a = Z_{11} I_{P1} + Z_{21} I_{P2}$$
$$0 = Z_{12} I_{P1} + Z_{22} I_{P2}.$$

In the second state,

$$0 = Z_{11} I_{P2} + Z_{21} I_{P2}$$
$$\mathcal{E}_a = Z_{12} I_{P2} + Z_{22} I_{P2}'.$$

To be consistent, the following condition must obtain:

$$\begin{bmatrix} Z_{11} & Z_{21} & 0 & \mathcal{E}_a \\ Z_{12} & Z_{22} & 0 & 0 \\ 0 & Z_{11} & Z_{21} & 0 \\ 0 & Z_{12} & Z_{22} & \mathcal{E}_a \end{bmatrix} = 0,$$

that is to say,

$$(Z_{21} - Z_{12})(Z_{11} Z_{22} - Z_{12} Z_{21}) = 0.$$

As we have seen above that $Z_{11}Z_{22} - Z_{12}Z_{21} \neq 0$, it is thus demonstrated that $Z_{11} - Z_{12} = 0$.

The extension to n coupled circuits instead of 2 would result from a similar mathematical procedure.

IV. Application to Coupled Circuits

A. Electric Quadrupole

Let electromotive forces $\mathcal{E}_{P1}$ and $\mathcal{E}_{P2}$ be applied on the two circuits simultaneously. The following equations are obtained

$$\mathcal{E}_{P1} = Z_{11} I_{P1} + Z_{21} I_{P2}$$
$$\mathcal{E}_{P2} = Z_{12} I_{P1} + Z_{22} I_{P2}.$$

Alternately $\mathcal{E}_{P2}$ and I_{P2} can be expressed in terms of $\mathcal{E}_{P1}$ and I_{P2}.

$$\mathcal{E}_{P2} = b_{11} \mathcal{E}_{P1} + b_{12} I_{P1}$$
$$I_{P2} = b_{21} \mathcal{E}_{P1} + b_{22} I_{P1}.$$

The reciprocal property of the mutual impedances $Z_{12} = Z_{21}$ results in a necessary condition for the b coefficients. For any physical quadrupole, the matrix

$$\begin{matrix} b_{11} & b_{12} \\ b_{21} & b_{22} \end{matrix}$$

of the coefficients of the above equations must be such as to have its determinant:

$$\begin{vmatrix} b_{11} & b_{12} \\ b_{21} & b_{22} \end{vmatrix}$$

equal to -1.

B. Computation of Mutual Impedances

The expression of the generalized mutual impedances has been found above. It is

$$Z_{21} = Z_{12} = -\int_{L1} \frac{I_1^*}{I_{P1}^*} \frac{E_{21} dl}{I_{P2}} = -\int_{L2} \frac{I_2^*}{I_{P2}^*} \frac{E_{12} dl}{I_{P1}}.$$

Let us consider the case of nondistributed currents in two closed circuits. Then

$$Z_{21} = Z_{12} = -\int_{L1} \frac{E_{21} dl}{I_2} = -\int_{L1} \frac{E_{12} dl}{I_1}.$$

[4] W. Dällenbach, "Reciprocity theorem of the electromagnetic field," *Archiv für Elek.*, vol. 36, pp. 153–165; March 31, 1942.

[5] IRE Standards on Antennas—Methods of Testing, 1948.

E_{21} can be expressed in terms of a scalar potential ϕ_2 and a vector potential A_2.

$$E_{21} = -\operatorname{grad}\phi_2 - \mu\frac{\partial A_2}{\partial t}\cdot$$

The circulation of the gradient of ϕ_2 along the closed path *L1* does not make any contribution to the results. On the other hand, the expression of A_2 is as follows

$$A_2 = \int_{L2}\frac{I_2\exp\left(-j\frac{\omega}{c}r\right)}{r}dl_2,$$

where r is the distance from the element dl_2 to the element dl_1 considered, and c is the speed of light in the medium (ϵ, μ).

It follows that

$$Z_{21} = j\omega\mu\int_{L1}\int_{L2}\frac{\exp\left(-j\frac{\omega}{c}r\right)}{r}dl_1dl_2.$$

This expression obviously verifies the reciprocal property of mutual impedances. In case

$$\frac{\omega}{c}r = \frac{r}{\frac{\lambda}{2\pi}}$$

is very small everywhere compared to unity, the above expression is closely approximated by

$$Z_{21} = Z_{12} = j\omega\mu\int_{L1}\int_{L2}\frac{dl_1dl_2}{r}\cdot$$

The mutual induction coefficient

$$M = \mu\int_{L1}\int_{L2}\frac{dl_1dl_2}{r}$$

is thus computable by the well-known Neumann formula. This coefficient becomes complex however as soon as the above-mentioned condition does not hold, that is to say, when the radiation field is not negligible.

V. Application to Linear Antennas

A. Radiation Coupling

In the case when the two linear antennas are located at a distance very great compared with $\lambda/2\pi$ and their own lengths, the fields E_{12} and E_{21} reduce to the radiation fields and can be considered as practically constant along the whole length of each antenna. The expressions for the mutual impedances are therefore as follows:

$$Z_{12} = -\frac{E_{21}}{I_{P2}}\int_{L1}\frac{I_1^*}{I_{P1}^*}\cos(E_{21}dl)dl$$

$$Z_{21} = -\frac{E_{12}}{I_{P1}}\int_{L2}\frac{I_2^*}{I_{P2}^*}\cos(E_{12}dl)dl.$$

In case the antennas are rectilinear, the cosines become constant. Should the antenna lengths be the same and *P1* and *P2* being corresponding points so that the current distribution becomes identical for both antennas, then the reciprocal property $Z_{12}=Z_{21}$ results in the following equation:

$$\frac{E_{21}}{E_{12}} = \frac{I_{P1}}{I_{P2}}\cdot$$

Finally, for same current amplitudes, the reciprocity between mutual coefficients leads to $E_{21}=E_{12}$, that is, to the particular form given to it by Sommerfeld and Pfrang[6] in 1926.

B. Cases for Zero Mutual Impedances

The expression given above shows the different cases for which mutual impedances can be equal to zero. Consider

$$Z_{12} = Z_{21} = -\frac{E_{21}}{I_{P2}}\int_{L1}\frac{I_1^*}{I_{P1}^*}\cos(E_{21}\cdot dl)dl.$$

This expression is zero: (a) when $E_{21}=0$. that is, when antenna 1 is located at a zero field radial with respect to antenna 2; (b) when $\cos(E_{21}\cdot dl)=0$, that is, when antenna 1 is at a right angle to the field radiated by antenna 2; and (c) when

$$\int_{L1}\frac{I_1^*}{I_{P1}^*}\cos(E_{21}dl)dl = 0,$$

that is, when the current distribution induced in antenna 1 is such that the work done by the radiated field due to antenna 2 along the length of antenna 1 is zero.

When $Z_{12}=0$ for any one of the above causes, then Z_{21} is necessarily also equal to zero. It may be, however, that the causes which make Z_{12} and Z_{21} simultaneously equal to zero result from two different causes among the three listed above.[1]

C. Reciprocal Properties of Same Antenna Used as Transmitting and Receiving Element

Let an antenna be submitted to experimental measurement of its directive properties. In a first series of experiments, let it be used as a transmitting element. The measuring apparatus will include a receiving antenna, and the equations relating the impressed electromotive force and the induced currents will be as before:

$$\mathcal{E} = Z_{11}I_{P1} + Z_{21}I_{P2}$$

$$0 = Z_{12}I_{P1} + Z_{22}I_{P2}.$$

However, the reaction of the measuring apparatus on the antenna under test will be made negligible, and the current measured I_{P2} will be very approximately

[6] A. Sommerfeld and H. Pfrang, "The reciprocity theorem," *Jahr, drahtl. Telegr.*, vol. 26, p. 93; 1926. Also, vol. 37, pp. 167–169; 1931.

equal to

$$I_{P2} = -\mathcal{E}\frac{Z_{12}}{Z_{11}Z_{22}}.$$

Let the system be reversed and the antenna under test be used as a receiving element. A signal generator will apply an electromotive force $\mathcal{E}'$ on the same auxiliary antenna as before, and the equations will become

$$0 = Z_{11}I_{P1}' + Z_{21}I_{P2}'$$
$$\mathcal{E}' = Z_{12}I_{P1}' + Z_{22}I_{P2}'.$$

In this case, however, the term $Z_{12}I_{P1}'$ will be negligible in the second equation and the measured current will be given by

$$I_{P1}' = -\mathcal{E}'\frac{Z_{21}}{Z_{11}Z_{22}}.$$

Thus the radiation diagram and radiation resistance of the same antenna used as a transmitting or receiving element are identical. The previous equations point out that correct experimental results will be obtained, however, in that case only where reaction of the measured on the measuring system can be made negligible.

VI. Application to Wave Projectors

The reciprocity theorem can also be applied to wave projectors. In this case, however, the most useful form is that due to H. A. Lorentz. Let two systems *D1* and *D2* be considered with wave projectors *W1* and *W2* (Fig. 5).

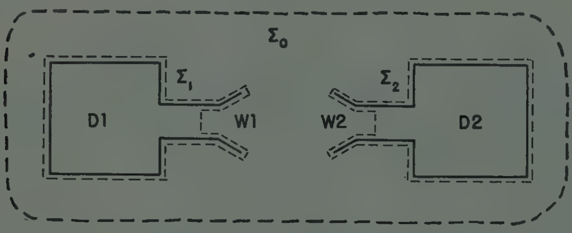


Fig. 5

Consider the space bounded by surfaces Σ_1 and Σ_2 on one side and an infinitely remote surface Σ_0 on the other side. Consider two possible electromagnetic states denoted by subscripts 1 and 2. The total flux of the vector product $\boldsymbol{E}_1 \times \boldsymbol{H}_2$ is equal to the total flux of the vector product $\boldsymbol{E}_2 \times \boldsymbol{H}_1$. As explained before, the flux through Σ_0, which can be assimilated to a sphere of infinite radius, is equal in both cases, so that the following equation[7] is obtained

$$\int_{\Sigma_1+\Sigma_2} (\boldsymbol{E}_1 \times \boldsymbol{H}_2)_n d\sigma = \int_{\Sigma_1+\Sigma_2} (\boldsymbol{E}_2 \times \boldsymbol{H}_1)_n d\sigma.$$

In the first state, let *W1* be a transmitter and *W2* a receiver. The fields in a cross section of the wave guide feeding *W1* will be denoted by $\boldsymbol{E}_1$, $\boldsymbol{H}_1$ and in a cross section of the feeder to *W2* by $\boldsymbol{e}_2$, $\boldsymbol{h}_2$. In the second case, the fields will be, respectively, $\boldsymbol{E}_2$, $\boldsymbol{H}_2$ and $\boldsymbol{e}_1$, $\boldsymbol{h}_1$. The above equation gives the following relation

$$\int_{\Sigma_1} (\boldsymbol{e}_2 \times \boldsymbol{H}_1)_n d\sigma + \int_{\Sigma_2} (\boldsymbol{E}_2 \times \boldsymbol{h}_1)_n d\sigma$$
$$= \int_{\Sigma_1} (\boldsymbol{E}_1 \times \boldsymbol{h}_2)_n d\sigma + \int_{\Sigma_2} (\boldsymbol{e}_2 \times \boldsymbol{H}_1)_n d\sigma.$$

Now, provided the field distribution across Σ_1 in the two cases are the same,

$$\int_{\Sigma_1} (\boldsymbol{E}_1 \times \boldsymbol{h}_2)_n d\sigma = -\int_{\Sigma_1} (\boldsymbol{h}_2 \times \boldsymbol{E}_1)_n d\sigma$$
$$= -\int_{\Sigma_1} (\boldsymbol{e}_2 \times \boldsymbol{H}_1)_n d\sigma,$$

and the same is true for the surface integral on Σ_2. Thus, the previous equation reduces to

$$\int_{\Sigma_1} (\boldsymbol{e}_2 \times \boldsymbol{H}_1) d\sigma = \int_{\Sigma_2} (\boldsymbol{e}_1 \times \boldsymbol{H}_2) d\sigma,$$

or again to

$$\int_{\Sigma_1} (\boldsymbol{e}_2 \times \boldsymbol{E}_1) d\sigma = \int_{\Sigma_2} (\boldsymbol{e}_1 \times \boldsymbol{E}_2) d\sigma.$$

This must be true at all times. Consider the case when E_1 and E_2 are sinusoidal and in phase:

$$E_1 = |E_1| \cos \omega t$$
$$E_2 = |E_2| \cos \omega t.$$

Let ψ_1 and ψ_2 be the phase shifts of e_1 and e_2 with respect to E_2 and E_1. Finally, let *W1* and *W2* be the transmitted and w_1 and w_2, the received powers through Σ_1 and Σ_2, respectively. The reciprocity theorem gives finally $(W_1w_2)^{1/2} \cos \omega t \cos (\omega t + \psi_2) = (W_2w_1)^{1/2} \cos \omega t \cos (\omega t + \psi_1)$.

This means that the time taken by the wave to travel from *D1* to *D2* in the first electromagnetic case is equal to the time taken from *D2* to *D1* in the second case, whatever may be the obstacles in the path. It also means that the ratio of transmitted to received power is the same in both cases.[7]

VII. Conditions for Validity of the Reciprocity Theorem

It should be emphasized, however, that all the above conclusions depend on the validity of Maxwell's equations at all points of the system considered. That is to say, the characteristic coefficients ϵ, μ, and γ should be independent of time and of electric and magnetic field strengths. More generally, the properties of the medium should be such that the characteristic tensors ϵ_{ik}, μ_{ik}, γ_{ik} should be independent of time and field strengths and, furthermore, symmetrical. It is from this latter condition that the reciprocal properties found above are derived.

In practice, such cases as ferromagnetic substances, electronic space charges, and ionized gases are excluded from the foregoing analysis.[2,4]

[7] H. Gutton and T. Ortusi, "Sur le théoréme de réciprocité en ondes hertziennes," *Compt. Rend. Acad. Sci.* (Paris), vol. 217, pp. 677–679; December 27, 1943.

Design Factors in Low-Noise Figure Input Circuits*

MATTHEW T. LEBENBAUM†, SENIOR MEMBER, IRE

Summary—**The use of double-tuned circuits in the inputs of high-frequency amplifiers often results in achieving better noise figures than if the simpler single-tuned network were used. This paper describes a method for calculating the minimum noise figure attainable with double-tuned circuits and provides simple data for designing the network. The results of the analysis, including both active and passive tube input loading, are presented in the form of a nomogram. The constants of a double-tuned input circuit that will provide minimum noise figure may be determined rapidly.**

Introduction

THE NOISE FIGURE of a receiver is largely determined in the low-level circuits of the receiver, and it is important that the signal circuit at these levels be carefully designed from the standpoint of noise if the minimum noise figure is to be obtained. This is especially true in microwave receivers where the signal is converted from rf to if without amplification (in fact, usually with a loss) so that any noise sources in the receiver before if amplification directly affect the noise figure.

I. Double-Tuned Circuit Problem

One serious source of noise occurs in the grid circuit of the first rf amplifier or (in microwave receivers) the first tube of the if amplifier. In the latter case, optimum matching networks can be realized rather easily since they do not have to be tunable. Herold[1] has analyzed the conditions for obtaining minimum noise figure in terms of the optimum value of the antenna conductance presented by the matching network to the input grid. The simplest type of matching network is the single-tuned. He pointed out, however, that, in certain cases, a double-tuned network is superior to it with respect to noise figure, but his analysis is not easily adapted to determining the design parameters of such networks. If secondary loading may be neglected, the design of a double-tuned transformer is not too difficult, but when tube input losses begin to load the transformer secondary not only does the actual determination of the parameters of a double-tuned transformer become more difficult, but so does the determination of the bandwidth for which the minimum noise figure is obtained.

This paper derives an expression for the noise figure of an amplifier with a transitionally[2] coupled double-tuned transformer input circuit. It then presents a nomographic solution of this expression that permits a simple graphical determination of the optimum bandwidth of the circuit.

Any calculation concerning noise figure must involve a number of quantities that are not easily measured. Such quantities as input conductance of vacuum tubes, percentage of the input conductance that is electronic in nature, of the crystal mixer, if impedances and equivalent tube noise resistances are difficult to determine accurately. Values may be assigned to these quantities and a noise figure calculated, but the absolute accuracy of the answer can only be that of the original data. The usefulness of an analysis such as that carried out in this paper might reasonably be questioned if the final result desired was an exact numerical value of noise figure. If, however, one desires a circuit design that will result in the *minimum* noise figure, an analysis of this type is useful. One may rapidly obtain curves of noise figure as a function of bandwidth, from which the effect of varying tube parameters is readily apparent. It will be found that the bandwidth for minimum noise figure is not extremely sensitive to these parameters, and a design center is easily established.

In some cases, the optimum bandwidth may be less than the required receiver bandwidth, as in wide-band low-frequency amplifiers. Then the input circuit must be designed for the required bandwidth, and the degradation in noise figure accepted. However, at the high frequencies, the bandwidth of the input circuit for minimum noise may be considerably wider than the required receiver bandwidth. The method of determining this bandwidth is presented in this paper.

II. Problem Treated Analytically

The circuit and the noise sources that are considered are shown in Fig. 1. Here the signal source and its internal impedance are represented by a constant-voltage generator in series with a resistance R_1. A shunt capacitance C_1, connected across the generator, represents the capacitance that exists when working from a mixer. The input circuit to the tube is represented by a shunt resistance R_2 and a shunt capacitance C_2. The circuit connecting the signal source to the input grid of the tube is a double-tuned transitionally coupled transformer defined by the parameters λ_1, λ_2, and λ_3. The sources of noise that may arise in this type of circuit are: (1) the thermal noise in the source resistance, (2) the thermal

* Decimal classification: R161.6×R361.211. Original manuscript received by the Institute, January 31, 1949; revised manuscript received, July 13, 1949. Presented, 1948 IRE National Convention, New York, N. Y., March 25, 1948.

† Airborne Instruments Laboratory, Inc., Mineola, L. I., N. Y.

[1] E. W. Herold, "An analysis of the signal-to-noise ratio of ultra-high-frequency receivers," *RCA Rev.*, vol. 6, pp. 302–331; January, 1942.

[2] Transitional coupling, sometimes called "flat-flat coupling," is that coupling which results in maximal flatness of the transfer impedance Z_{12} as a function of frequency. Mathematically, transitional coupling occurs when

$$\frac{dz_{12}}{df} = \frac{d^2z_{12}}{df^2} = \frac{d^3z_{12}}{df^3} = 0.$$

The frequency f_0 where this condition exists, is defined as the center frequency.

noise produced by the passive loading—the ohmic portion of the transformer and tube losses, (3) the noise produced by the active tube loading—transit-time loading, and (4) shot and partition effect in the tube itself. These noise sources are shown in Fig. 1. The circuit is in

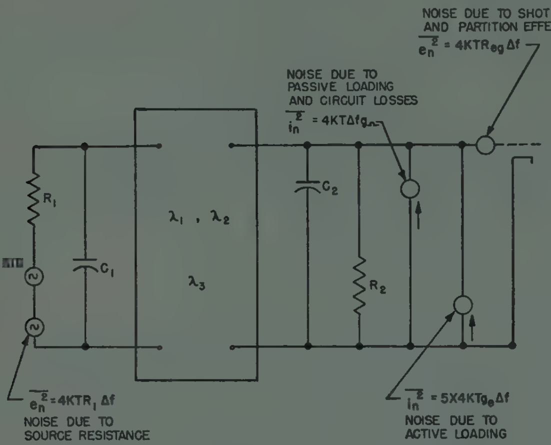


Fig. 1—Equivalent input circuit including sources of noise.

the same form as that used by Herold[1] except for the network coupling the signal source to the input grid. The input loading has been broken into two components—one at the ambient temperature (representing the ohmic losses) and one at five times the ambient temperature (representing the electronic loading caused by the transit-time effect).[3]

In the derivation of the expression for noise figure, the following terminology will be used

K = Boltzmann's constant = 1.37×10^{-23} joules per degree Kelvin
T = temperature, degrees Kelvin
R_1 = source resistance

$$R_2 = \frac{1}{g_\Omega + g_e} = \text{total secondary loading}$$

g_Ω = conductance due to passive tube loading and circuit losses
g_e = conductance due to electronic or transit-time loading
R_{eq} = tube equivalent noise resistance
Δf = frequency band over which noise is measured and
$\lambda_1, \lambda_2, \lambda_3$ = transformer parameters.

From Fig. 1, the values of the various noise components referred to the input grid may be written. The thermal noise due to R_1, is given by

$$\overline{e_n^2} = \left(\frac{4KT\Delta f}{R_1}\right) Z_T^2 \tag{1}$$

where Z_T is the transfer impedance of the network.

The noise due to the input conductance of the tube, both active and passive, is given by

$$\overline{e_n^2} = 4KT\Delta f\left(\frac{1}{R_2} + 4g_e\right) Z_2^2 \tag{2}$$

where Z_2 is the secondary driving-point impedance of the network.

The noise due to shot and partition effect is given by

$$\overline{e_n^2} = 4KTR_{eq}\Delta f \tag{3}$$

where R_{eq}, the equivalent tube noise resistance, is defined as that resistance which, if placed between grid and ground of a noise-free tube, would produce in the plate current a noise current equal to that which the actual "noisy" tube produces with its grid shorted to ground.

The summation of (1), (2), and (3) represents the square of the total noise voltage appearing at the grid of the first tube. Thus

$$\overline{e_{nt}^2} = 4KT\Delta f\left(\frac{Z_T^2}{R_1} + \frac{1 + 4g_eR_2}{R_2} Z_2^2 + R_{eq}\right). \tag{4}$$

The square of the signal voltage referred to the input grid is

$$e_s^2 = \left(\frac{e_{sig}}{R_1} Z_T\right)^2. \tag{5}$$

The signal-to-noise power ratio is obtained by dividing (4) by (5)

$$\frac{e_s^2}{e_{nt}^2} = \frac{e_{sig}^2}{4KTR_1\Delta f}\left[\frac{Z_T^2}{R_1\left(\frac{Z_T^2}{R_1} + \frac{1+4g_eR_2}{R_2} Z_2^2 + R_{eq}\right)}\right]. \tag{6}$$

The quantity in front of the brackets may be recognized as the signal-to-noise ratio of an ideal receiver; that is, a receiver in which the only noise output is that introduced by the thermal noise of the source resistance. The expression within the brackets is a quantity that represents the degradation of the signal-to-noise ratio caused by noise sources in the input circuit and the input tube; the reciprocal of this quantity is the noise figure. It is then possible, by utilizing the excellent material contained in a report by Stone and Lawson,[4] to reduce the expression, as done in the Appendix, within the brackets to the following rather simple expression for the noise figure[5]

$$F = 1 + A(1 + 4g_eR_2) + BR_{eq}C_2\omega_0 \tag{7}$$

where

$$A = \frac{\alpha_2}{\alpha_1} \frac{(\frac{1}{8}\alpha_2^2 + \frac{3}{8}\alpha_1^2)^2 + \alpha_1^2}{\frac{1}{2}(\alpha_1^2 + \alpha_2^2) + \frac{1}{64}(\alpha_1^2 + 3\alpha_2^2)(3\alpha_1^2 + \alpha_2^2)}$$

$$B = \frac{1}{\alpha_1} \frac{\alpha^4\left(\frac{1}{4} + \frac{\alpha^2}{64}\right)}{\frac{1}{2}(\alpha_1^2 + \alpha_2^2) + \frac{1}{64}(\alpha_1^2 + 3\alpha_2^2)(3\alpha_1^2 + \alpha_2^2)}$$

[3] D. O. North and W. R. Ferris, "Fluctuations induced in vacuum tube grids at high frequencies," Proc, I.R.E., vol. 29, pp. 49–50; February, 1941.

[4] J. L. Lawson and A. M. Stone, "The double-tuned circuit with transitional coupling," Radiation Laboratory Report No. 784, Massachusetts Institute of Technology, Cambridge, Mass.

[5] H. T. Friis, "Noise figures of radio receivers," Proc. I.R.E., vol. 32, pp. 419–422; July, 1944.

$$\alpha_1 = \frac{1}{R_1C_1\omega_0} = \text{primary dissipation factor}$$

$$\alpha_2 = \frac{1}{R_2C_2\omega_0} = \text{secondary dissipation factor}$$

$$\alpha = \alpha_1 + \alpha_2.$$

This is the equation upon which the nomogram is based. The quantities involved are readily determinable or easily calculable from known quantities of the tube and the signal source. For example, the quantity g_eR_2 is the ratio of the electronic input conductance of the tube to the total input conductance. R_{eq} is the equivalent noise resistance of the tube. f_0 is the center frequency of the matching network. C_2 is the input capacitance of the tube. The quantities A and B are both functions *only* of the primary and secondary dissipation factors α_1 and α_2.

III. Nomographic Solution

Fig. 2 is a nomographic solution of (7). From it, the noise figure obtainable from a given tube and matching network may be readily determined. In Chart I is plotted a family of curves with the secondary dissipation factor α_2 as the abscissa and the first two quantities of the noise-figure expression as the ordinate. The family parameter b is the fractional bandwidth (the ratio of input-circuit bandwidth to center frequency). α_2 is a fundamental parameter of the tube and the center frequency that is chosen, R_2 being the input resistance, and C_2 the input capacitance. Chart II is a family of curves in which the ordinate B is from the third term of (7), and in which the family parameter is the same b as that used in Chart I. Chart III is simply a mechanism for multiplying B by $R_{eq}C_2\omega_0$, which is again a function of the tube alone. In Chart IV, are plotted curves of constant noise figure derived from the summation of the two factors obtained from Charts I and III.

The method of using the nomogram is simple and straightforward. Entering the α_2 axis, at a point determined by the tube parameters, one draws a vertical line intersecting an arbitrarily selected b curve. A quantity selected from the scale in the upper left-hand corner of Chart I (which again is a function of the tube and represents that percentage of the input conductance contributed by the electronic loading) is added vertically to the previous intersection and a horizontal line drawn into Chart IV. Returning to the α_2 axis, the vertical line

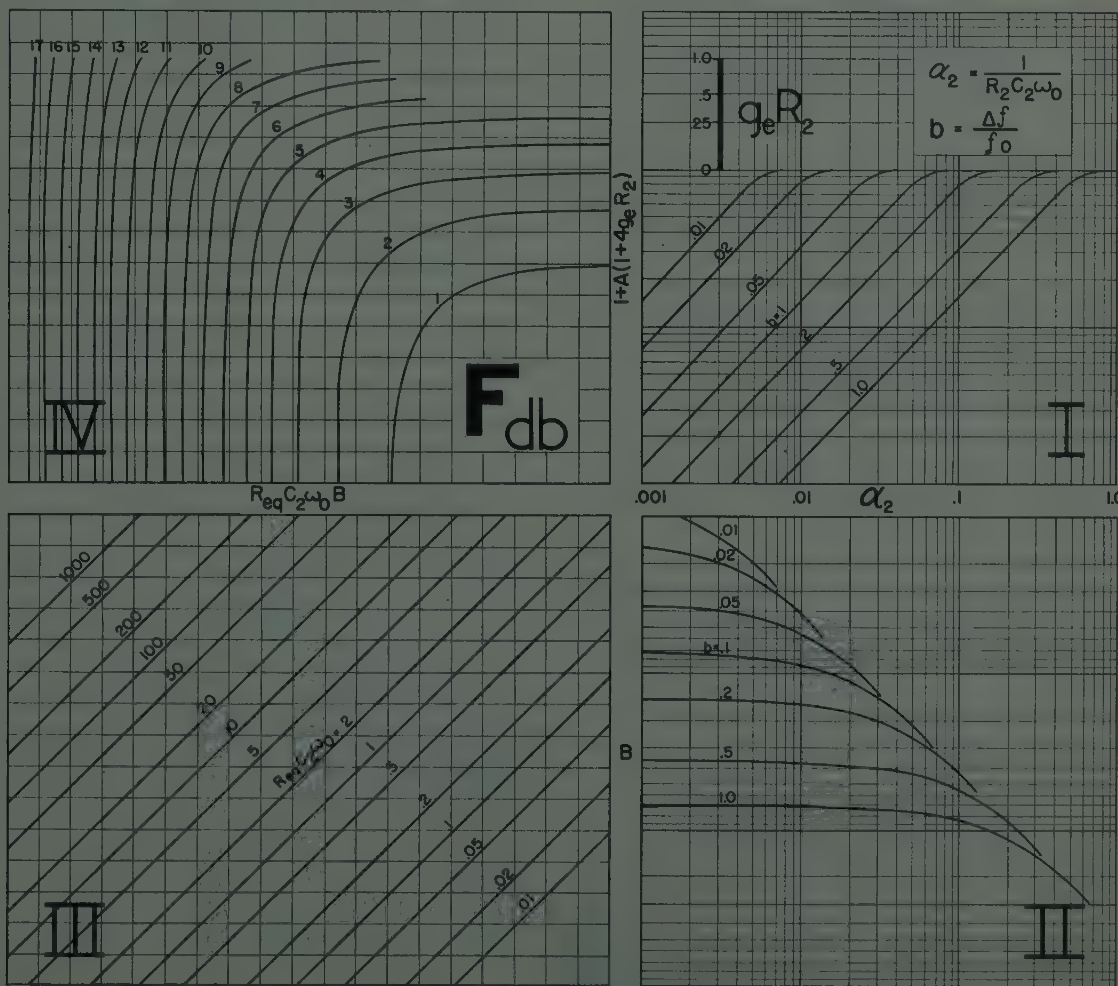


Fig. 2—Nomographic solution of the noise-figure equation (7).

is drawn downward into Chart II, intersecting the same b curve, thence horizontally into Chart III to an intersection with the proper multiplying factor, and vertically into Chart IV to an intersection with the first line. This intersection then gives a noise figure for the particular bandwidth chosen for the input circuit.

It is simple to determine the noise figure for a number of selected input bandwidths and to draw a curve of noise figure versus bandwidth. For example, if an amplifier having a bandwidth of 10 Mc and centered at 160 Mc is to be built, using pentode-connected 6AK5's, the following parameters may be assumed: an input resistance of 2,000 Ω, a noise equivalent resistance of 2,000 Ω, and an input capacitance of 7 μμf. α_2 is then 0.07. It is further assumed that 50 per cent of the input loading is electronic in nature. Fig. 3 shows a plot of the

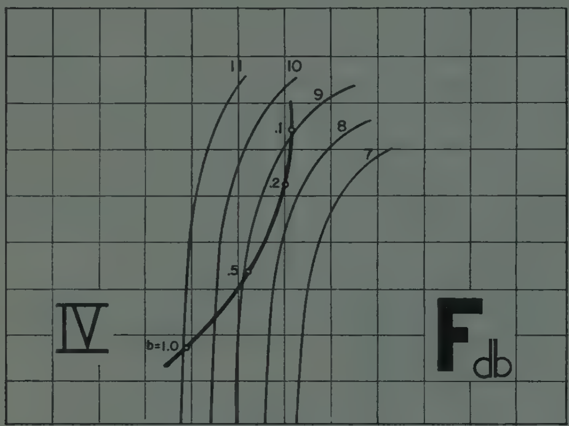


Fig. 3—Effect on noise-figure F of varying the fractional bandwidth b through values 0.1, 0.2, 0.5, and 1.0.

over-all receiver noise figure when the input bandwidth factor b is 0.1, 0.2, etc. This corresponds to bandwidths of 16 Mc, 32 Mc, etc. It may be seen that a minimum occurs somewhere between 0.25 and 0.3 (between 40 Mc and 52 Mc); the actual value is not too critical. The noise figure is not a sensitive function of bandwidth in the region of the minimum and changing the assumed values appreciably does not change the value of F significantly.

Returning to the nomogram, one may see that it permits a convenient means of visualizing how changes of parameters effect the noise figure. For instance, if one is working in Chart IV in a region where the lines are practically vertical, it is evident that varying $R_{eq}C_2\omega_0B$ will have a much greater effect than varying the other factor; in the region of horizontal contours, the reverse is true.

IV. Reliability of Results

It must be repeated that the method described gives results whose accuracy will depend on the basic assumptions made. Two classes of error are possible. The first class is related to the errors made in choosing values for the basic parameters. The effect of errors of the first class can be checked easily by repeating the calculations over the range of expected variations.

The second class of errors deserves more careful attention. In the preparation of the nomogram, it has been assumed that the secondary driving-point impedance Z_2 of the input circuit remains constant and equal to its value at the center frequency. So long as α_1 is equal to or greater than α_2, this is quite accurate. Another assumption made is that the value of R_2 should remain constant throughout the pass band. This is usually not true because the input conductance of a vacuum tube varies as the square of the frequency; with fractional bandwidths of the order of 0.1 or more, it may appear at first that the errors would be substantial. It has also been assumed that both Z_2 and Z_T have small phase angles over most of the pass band. This is also quite accurate if $\alpha_1 \geqq \alpha_2$.

Careful use of the nomogram, however, will show that semiquantitative information is actually available despite the fact that some of the assumptions made appear inconsistent with reality. This fact can best be explained by means of an example. Suppose that the previous problem were to be solved again. An amplifier is desired with a 10-Mc bandwidth, centered at 160 Mc. It has already been determined that the optimum figure is obtained for a fractional bandwidth between 0.25 and 0.3. This corresponds to a change in the value of R_2 between extremes of the band by a factor of 1.7. If one takes for R_2 values corresponding to the low-frequency edge and the high-frequency edge of this bandwidth and recalculates the optimum bandwidth, it will be found that the arithmetical average gives a resultant optimum bandwidth of 0.27. This value is the same as that determined by simply taking the value of R_2 at the center frequency. Obviously, this result is not quite accurate because the circuit will have transitional coupling only at the center frequency and will deviate from this condition at the edges of the band. This effect is small in all practical cases.

It is most important to recognize that the calculation of noise figure is exact only at the center frequency. To be strictly correct, one should integrate the noise figure over the entire bandwidth. The results obtained from a single-frequency noise calculation are always on the optimistic side because the noise figure deteriorates toward the edges of the pass band. This point must be considered when extremely large over-all bandwidths are required, if certain types of circuits (such as grounded-grid stages) are employed. The noise figure of grounded-grid stages deteriorates considerably toward the edges of the pass band; therefore, noise figures calculated on a single-frequency basis are likely to be very seriously in error if this type of tube is used with an input circuit whose bandwidth approaches that of the over-all amplifier bandwidth. For example, in the Wallman-cascode[6] circuit, it has been noted that serious degradation of the over-all noise figure is suffered unless the bandwidth of the input to the grounded-grid portion

[6] H. Wallman, A. B. MacNee, and C. P. Gadsden, "A low-noise amplifier," Proc. I.R.E., vol. 36, pp. 700–708; June, 1948.

of the cascode is very much wider than the over-all receiver bandwidth.

It should be also noted that no account has been taken of the possible improvement that may be obtained at high frequencies by suitably mistuning the input circuit.[7] This is difficult to treat qualitatively and is usually achieved by purely empirical adjustments. In any event, this improvement can be attained only in the narrow-band case, and this paper treats only the wide-band.

Once the desired bandwidth has been determined from the nomogram, the calculation of the actual parameters of the transformer is a relatively simple matter. A résumé of the applicable material in the Stone and Lawson[4] report will be given here.

Table I lists the known parameters; b is the bandwidth (determined from the nomogram) that gives the minimum noise figure; the other quantities are self-explanatory. α_2 is known from the tube parameters; one may then determine α_1 (and therefore C_1, the capacitance which the input circuit must present). If one is fortunate, this quantity is greater than the mixer capacitance. If it is not, the ideal bandwidth may not be realized, and the closest bandwidth for which C_1 is physically realizable must be chosen. From the values of α_1, α_2, C_2, and ω_0, all the constants of the transformer may be calculated from Table I.

TABLE I

PROCEDURE FOR CALCULATING THE TRANSFORMER CONSTANTS

Given:	$b = \dfrac{\Delta f}{f}$ (from nomogram)
	C_2 = tube input capacitance
	R_2 = tube input loading and circuit losses
	R_1 = source resistance
	$\omega_0 = 2\pi f_0$ = radian center frequency
Calculate:	$\alpha_2 = \dfrac{1}{R_2 C_2 \omega_0}$
	$\alpha_1 = \sqrt{2} f - \alpha_2$
	$C_1 = \dfrac{1}{R_1 \omega_0 \alpha_1}$
Transformer constants:	
	$\lambda_3 = \dfrac{1}{C_1 C_2 \omega_0^4}$
	$\lambda_2 = \dfrac{1}{C_2 \omega_0^2}\left(1 + \dfrac{\alpha_2^2}{8} + \dfrac{3\alpha_1^2}{8}\right)$
	$\lambda_1 = \dfrac{1}{C_1 \omega_0^2}\left(1 + \dfrac{\alpha_1^2}{8} + \dfrac{3\alpha_2^2}{8}\right)$
	$M = (\lambda_1 \lambda_2 - \lambda_3)^{1/2}$

There are various ways in which these constants may be realized in physical transformers. Fig. 4 shows three possible arrangements. The mutually coupled circuit is directly related to λ_1, λ_2, and a value of M, the mutual coupling that may be calculated from the values of the λ's (as given in Table I). In certain cases, the mutually coupled transformer may be transformed into π or T sections, which are often more easily constructed at the higher frequencies, especially for very wide bandwidths. Table II shows the transformation equations to convert from double-tuned to either of the other two transformers.

[7] M. J. O. Strutt and A. van der Zeil, "Signal-to-noise ratio at vhf," *Wireless Eng.*, vol. 23, pp. 241–249; September, 1946.

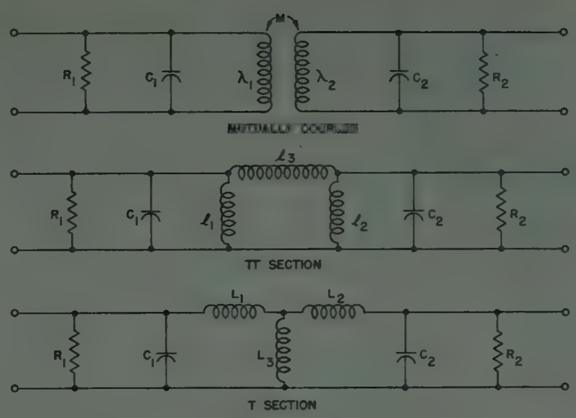


Fig. 4—Three common forms of the double-tuned circuit.

TABLE II

EQUIVALENCES BETWEEN FORMS OF THE DOUBLE-TUNED CIRCUIT

Doubled-tuned to T	$L_1 + L_3 = \lambda_1$	$\lambda_1 - M = L_1$
	$L_2 + L_3 = \lambda_2$	$\lambda_2 - M = L_2$
	$L_3 = M$	$M = L_3$
Double-tuned to π	$\dfrac{\lambda_1\lambda_2 - M^2}{\lambda_2 - M} = l_1$	$\dfrac{l_1 l_3 + l_1 l_2}{l_1 + l_2 + l_3} = \lambda_1$
	$\dfrac{\lambda_1\lambda_2 - M^2}{\lambda_1 - M} = l_2$	$\dfrac{l_2 l_3 + l_1 l_2}{l_1 + l_2 + l_3} = \lambda_2$
	$\dfrac{\lambda_1\lambda_2 - M^2}{M} = l_3$	$\dfrac{l_1 l_2}{l_1 + l_2 + l_3} = M$
T to π	$\dfrac{L_1L_2 + L_2L_3 + L_1L_3}{L_2} = l_1$	$\dfrac{l_1 l_3}{l_1 + l_2 + l_3} = L_1$
	$\dfrac{L_1L_2 + L_2L_3 + L_1L_3}{L_1} = l_2$	$\dfrac{l_2 l_3}{l_1 + l_2 + l_3} = L_2$
	$\dfrac{L_1L_2 + L_2L_3 + L_1L_3}{L_3} = l_3$	$\dfrac{l_1 l_2}{l_1 + l_2 + l_3} = L_3$
$\lambda_3 = L_1L_2 + L_2L_3 + L_1L_3$		

V. ACKNOWLEDGMENT

The author wishes to express his appreciation to E. G. Fubini and F. C. Clement of the Airborne Instruments Laboratory—the former for his suggestions concerning the construction of the nomogram from (7), and the latter for performing most of the numerical calculations.

APPENDIX

Derivation of Equation (7)

The signal-to-noise ratio has been shown to be

$$\frac{e_s^2}{e_{nt}^2}=\frac{e_{sig}^2}{4KTR_1\Delta f}\left[\frac{Z_T^2}{R_1\left(\frac{Z_T^2}{R_1}+\frac{1+4g_eR_2}{R_2}Z_2^2+R_{eq}\right)}\right]. \quad (6)$$

The reciprocal of the bracketed expression in (6) is the desired noise figure, F

$$F = R_1\frac{\left(\frac{Z_T^2}{R_1}+\frac{1+4g_eR_2}{R_2}Z_2^2+R_{eq}\right)}{Z_T^2}. \quad (8)$$

Simplifying (8), we obtain

$$F = 1+\frac{R_1}{Z_T^2}\left(\frac{Z_2^2}{R_2}(1+4g_eR_2)+R_{eq}\right). \quad (9)$$

From Stone and Lawson,[4] we obtain expressions for $|Z_2|^2$ and $|Z_T|^2$

$$|Z_2|^2=\frac{1}{C_2^2\omega_0^2}\;\frac{(\frac{1}{8}\alpha_2^2+\frac{3}{8}\alpha_1^2)^2+\alpha_1^2}{\alpha^4\left(\frac{1}{4}+\frac{\alpha^2}{64}\right)} \text{ at } \omega=\omega_0 \quad (10)$$

and

$$|Z_T|^2=\frac{1}{C_1C_2\omega_0^2}\;\frac{\frac{1}{2}(\alpha_1^2+\alpha_2^2)+\frac{1}{64}(\alpha_1^2+3\alpha_2^2)(3\alpha_1^2+\alpha_2^2)}{\alpha^4\left(\frac{1}{4}+\frac{\alpha^2}{64}\right)} \text{ at } \omega=\omega_0. \quad (11)$$

Dividing (10) by (11) and multiplying both sides by ω_0R_1/ω_0R_2 we obtain

$$\frac{R_1|Z_2|^2}{R_2|Z_T|^2}=\frac{\omega_0R_1C_1}{\omega_0R_2C_2}\;\frac{(\frac{1}{8}\alpha_2^2+\frac{3}{8}\alpha_1^2)+\alpha_1^2}{\frac{1}{2}(\alpha_1^2+\alpha_2^2)+\frac{1}{64}(\alpha_1^2+3\alpha_2^2)(3\alpha_1^2+\alpha_2^2)}$$

$$=\frac{\alpha_2}{\alpha_1}\;\frac{(\frac{1}{8}\alpha_2^2+\frac{3}{8}\alpha_1^2)^2+\alpha_1^2}{\frac{1}{2}(\alpha_1^2+\alpha_2^2)+\frac{1}{64}(\alpha_1^2+3\alpha_2^2)(3\alpha_1^2+\alpha_2^2)}=A. \quad (12)$$

Taking the reciprocal of (11) and multiplying by R_1/ω_0C_2 we obtain

$$\frac{R_1}{\omega_0C_2|Z_T|^2}=\frac{1}{\alpha_1}\;\frac{\alpha^4\left(\frac{1}{4}+\frac{\alpha^2}{64}\right)}{\frac{1}{2}(\alpha_1^2+\alpha_2^2)+\frac{1}{64}(\alpha_1^2+3\alpha_2^2)(3\alpha_1^2+\alpha_2^2)}=B. \quad (13)$$

Substituting (12) and (13) into (8) we obtain

$$F = 1+A(1+4g_eR_2)+BR_{eq}C_2\omega_0 \quad (7)$$

where A and B are as above.

On the Deduction of the Refractive Index Profile of a Stratified Atmosphere from Radio Field-Strength Measurements

JAMES W. GREEN†, ASSOCIATE IRE

Summary—**A method developed by Macfarlane[1] for the theoretical deduction of the refractive index profile of a horizontally stratified atmosphere from radio field-strength measurements was applied to several sets of radio field-strength data for which complementary measured refractive index profiles were available, for both standard and nonstandard meteorological conditions.**

The formal theoretical validity of Macfarlane's method was affirmed for the case of a linear *M* curve, $dM/dH=3.6$ units/100 feet, by application of the method to several theoretical radio field-strength profiles given by Standard Diffraction Theory. *M* curves were derived which were linear and possessed slopes within 3 per cent of the control slope of 3.6 *M* units/100 feet However, application of the method to the theoretical radio field-strength profiles of Standard Diffraction Theory revealed that in order to yield satisfactory refractive index data, a higher degree of accuracy was required of the radio field-strength data than one could expect to realize experimentally. The inherent sensitivity of the method to small, inevitable inaccuracies in the radio field-strength data makes it impossible to apply the method directly to experimental data. It is often possible, however, to approximate a radio field-strength profile, within the limits of experimental error, by an analytical curve or a combination of curves appropriately joined together. Applying this technique to experimental radio field-strength profiles for radio frequencies between 100 and 1,000 Mc, recorded under essentially standard meteorological conditions, it was found that Macfarlane's method yielded refractive index profiles which were in satisfactory agreement with the average experimentally measured refractive index profile.

In the important and most interesting case of nonstandard propagation, the method fails to provide refractive index profiles which are reliable or in consistent quantitative agreement with average experimentally measured profiles.

In experience, the simplifying assumptions of a horizontally stratified atmosphere and single mode propagation, essential to the theoretical formulation of the method, are, at best, only approximately fulfilled. However, even in those cases where the conditions of hori-

* Decimal classification: R271×R113.230.9. Original manuscript received by the Institute, March 1, 1949; revised manuscript received, August 30, 1949. Presented, joint meeting of the International Scientific Radio Union, and The Institute of Radio Engineers, Washington, D. C. on October 8, 1948.

The above entitled paper contains opinions of the author and should not be construed as official or reflecting the views of the Navy Department or the Naval Service at large.

† U. S. Navy Electronics Laboratory, San Diego, Calif.

[1] G. G. Macfarlane, "A method for deducing the refractive-index profile of a stratified atmosphere from radio observations," Meteorological Factors in Radio-Wave Propagation, Report of a Conference Held on April 8, 1946, at the Royal Institution, London, by the Physical Society and Royal Meteorological Society, published by the Physical Society, London.

zontal stratification and single mode propagation are fulfilled to a relatively good degree of approximation, and hence the recorded radio field-strength amplitudes themselves are relatively good approximations to those which would be expected ideally, the method yields refractive index profiles which are, in general, unsatisfacory approximations to the prevailing average measured profiles.

Because of the failure of the assumptions of horizontal stratification and single mode propagation to be fulfilled even approximately in most cases in experience, and because of the failure of the Macfarlane method to provide, in general, a good approximate answer in cases where these conditions are approximately fulfilled, it is concluded that the method is inadequate for the description of experimentally observed propagation phenomena.

Introduction

A METHOD IS presented by Macfarlane[1] whereby the refractive index profile of a stratified atmosphere may be derived theoretically from observations of the radio field-strength profile. Observations of amplitude only, and not of phase, of the radio field are required. Macfarlane further suggests that the refractive index profile derived solely from the radio observations may well represent a better average for use in predicting field strengths on different wavelengths than is obtained from a single meteorological sounding. The method depends on the assumption that the atmosphere is horizontally stratified and that, in consequence, the electromagnetic field may be represented as a sum of modes, and further that at distances "sufficiently far" beyond the optical horizon from a radio transmitter only one mode is of significance.

The equation of tropospheric wave propagation in a horizontally stratified atmosphere is:

$$\partial^2 W/\partial h^2 + \partial^2 W/\partial x^2 + k^2 M^2(h) W = 0 \tag{1}$$

where $M(h) \equiv n(h) + (h/R_0)$. $M(h)$ is the modified refractive index, $k = (2\pi/\lambda)$, $n(h)$ is the index of refraction, R_0 is the earth's radius, and h (also H) is the elevation above the earth's surface. A solution of (1) is

$$W = \sum_{n=1}^{\infty} \alpha_n U_n(h) \exp(-ikx \cos \theta_n). \tag{2}$$

$U_n(h)$ is a solution of the one-dimensional equation:

$$d^2 U_n/dh^2 + k^2(M^2(h) - \cos^2 \theta_n) U_n = 0. \tag{3}$$

Now, at distances "sufficiently far" beyond the horizon from a radio transmitter it is generally assumed that only the first term of the series of (2) remains of significant amplitude. Macfarlane shows that it is then possible to express $M(h)$ as a function of the observed quantity $A(h)$, the field strength, and to eliminate the phase, ϕ, which is difficult to measure. The assumed solution of (3) is then:

$$U(h) = A(h) \exp(i\phi). \tag{4}$$

Putting $\cos^2 \theta = a + ib$, substituting (4) in (3) and eliminating ϕ, the following expression results:

$$(a - M^2) = \frac{1}{k^2 A}\frac{d^2 A}{dh^2} - \frac{k^2 b^2}{A^4}\left\{\int_0^H A^2 dh\right\}^2. \tag{5}$$

Since $a = 1 + \epsilon$, where ϵ is a small constant independent of height, (5) may be written approximately as follows:

$$(M - 1 - \epsilon/2) = \frac{1}{2}\left\{\frac{k^2 b^2}{A^4}\left[\int_0^H A^2 dh\right]^2 - \frac{1}{k^2 A}\frac{d^2 A}{dh^2}\right\}. \tag{6}$$

Thus $M(h)$ is given, apart from a small constant which is independent of height. For convenience, let

$$\widehat{M} \equiv (M - 1 - \epsilon/2)10^6.$$

Macfarlane suggests two ways in which $\widehat{M}(h)$ may be found by measurement of field strength: (A) From one set of radio field-strength measurements at a fixed range giving values of $A(h)$ and a few observations of field strength at constant height and different ranges giving $kb/2$, the attenuation constant; (B) from two sets of radio field-strength measurements, each at a different wavelength.

Employing Macfarlane's method (A), refractive index profiles were deduced from radio field-strength profiles for the following cases:

(I) Theoretical radio field-strength profiles[2,3,4] given by Standard Diffraction Theory for a standard atmosphere ($K = 1.33$) for frequencies of 65, 170, 520, and 1,000 Mc.

(II) Theoretical radio-field strength profiles for 10-cm and 3-cm waves as derived by Pekeris[5] for a linear-exponential type M curve.

(III) Experimental overwater radio field-strength profiles for frequencies of 65, 170, and 520 Mc recorded under "standard" atmospheric conditions.

(IV) Experimental overwater radio field-strength profiles for 65 Mc recorded under nonstandard atmospheric conditions (an elevated subsidence inversion).

(V) Experimental overland radio field-strength profiles[6] for 520 Mc, recorded under standard and nonstandard atmospheric conditions.

I. Derivation of Modified Index of Refraction Profiles from Theoretical Radio Field-Strength Data Given by Standard Diffraction Theory

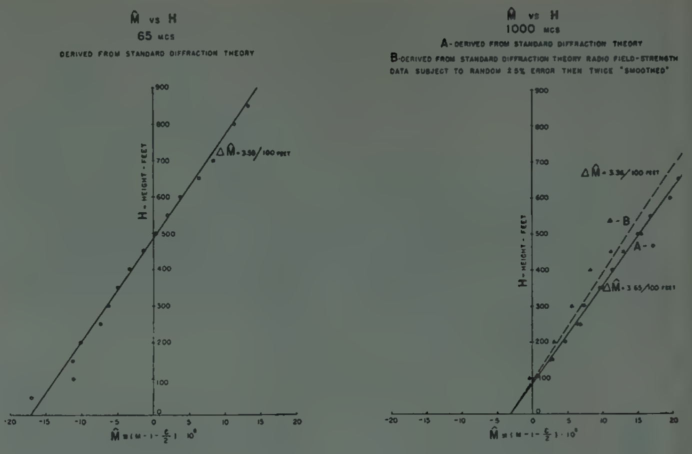
It was thought that it would be of academic interest to first apply Macfarlane's method to the theoretical radio height-gain profiles given by Standard Diffraction Theory where a relation between radio height-gain and meteorological profiles is established. Thus, in Standard Diffraction Theory, one assumes a linear modified

[2] G. N. Watson, "The diffraction of electric waves by the earth," *Proc. Roy. Soc.*, vol. 95; October, 1918 and July, 1919.
[3] K. A. Norton, "The calculation of ground-wave field intensity over a finitely conducting spherical earth," Proc. I.R.E., vol. 29, pp. 623–639; December, 1941.
[4] J. B. Smyth and M. D. Rocco, "Diffraction of high-frequency radio waves around the earth," Proc. I.R.E., vol. 37, pp. 1195–1203; October, 1949.
[5] C. L. Pekeris, "Wave theoretical interpretation of propagation of 10-cm and 3-cm waves in low level ocean ducts," Proc. I.R.E., pp. 453–462; May, 1947.
[6] J. P. Day and L. G. Trolese, "The propagation of short radio waves over desert terrain," to be published in Proc. I.R.E.

index profile for a standard atmosphere of $dM/dH = 0.036$ units/foot ($K=1.33$) and derives radio field-strength profiles for various frequencies.[1,2] If now one applies Macfarlane's method to the S.D.T.[7] radio field-strength profiles for several frequencies, one should be able to derive a modified refractive index profile which is linear and of slope $d\widehat{M}/dH = 0.036$ units/ft. Thus, one could establish the validity of the Macfarlane method for the case of a linear modified refractive index profile and also obtain knowledge as to the requisite relative degree of accuracy of the radio field-strength data. $\widehat{M}(h)$ profiles were derived by Macfarlane's method from S.D.T. radio field-strength data on 65, 170, 520 and 1,000 Mc for a distance of 100 statute miles, a transmitter height of 200 feet and horizontal polarization. The integral, $\int_0^H A^2 dh$, was computed by Simpson's Parabolic Rule. The expression $1/k^2A \; d^2A/dh^2$ was evaluated by the Method of Second Differences. It was found necessary to employ extreme accuracy and multiple "smoothing"[8] in the determination of the radio field-strength data from S.D.T. in order to insure that the derived $\widehat{M}$-profile data were reasonably consistent. As anticipated, extreme accuracy was found to be particularly important at the lower frequencies considered (i.e., 65 and 170 Mc) where it was found necessary to resort to exhaustive "smoothing" of the radio field-strength data. However, with this extreme degree of accuracy and repetitive "smoothing," $\widehat{M}$ curves were derived which were linear and possessed slopes within 3 per cent of the control slope of 3.6 μ units/100 ft. Figs. 1(a) and (b) show the derived $\widehat{M}$ curves for 65 and 1,000 Mc. To examine the effect of small errors in the radio field-strength data, the values of field strength, A, for 1,000 Mc were subjected to a random ± 5 per cent error (which is equivalent to about ± 0.5 db error). The resulting data was then twice "smoothed" and the $\widehat{M}$ curve derived (dotted line, Fig. 1(b)). A "best line" through the points had a slope of $(\Delta\widehat{M}/\Delta H) = 3.36$ units/100 feet and hence differed 7 per cent from the slope of the previously derived $\widehat{M}$ curve, with a maximum scatter of ± 2 $\widehat{M}$ units between points. Of the four frequencies considered, the computations for the highest frequency, 1,000 Mc, would be *least affected* by errors in the radio field-strength data (see Discussion) and hence would yield the *most favorable* agreement with the previously derived $\widehat{M}$ curve. A closer numerical examination of this sensitivity revealed that in order to insure that the variations in the derived values of $\widehat{M}$ lie within ± 1 M unit, errors in the field strength "A" must be less than ± 5 per cent at 1,000 Mc, while at 65 Mc such errors would have to be less than ± 0.02 per cent! It is seen then, that the method is extremely sensitive to small errors in the radio field-strength data, and it was anticipated that difficulty would be experienced in applying the method to experimental radio field-strength data where the relative accuracy is of the order of ± 1.0 db.

[7] Hereafter the term "Standard Diffraction Theory" will be referred to as S.D.T. for brevity.

[8] A. G. Worthing and J. Geffner, "Treatment of Experimental Data," John Wiley and Co., Inc., New York, N. Y., 1943.



(a) (b)

Fig. 1—(a) Modified refractive index profiles, derived by applying MacFarlane's Method to theoretical radio field-strength profiles given by Standard Diffraction Theory.
(b) Modified refractive index profiles, derived by applying Macfarlane's Method to theoretical radio field-strength profiles given by Standard Diffraction Theory.

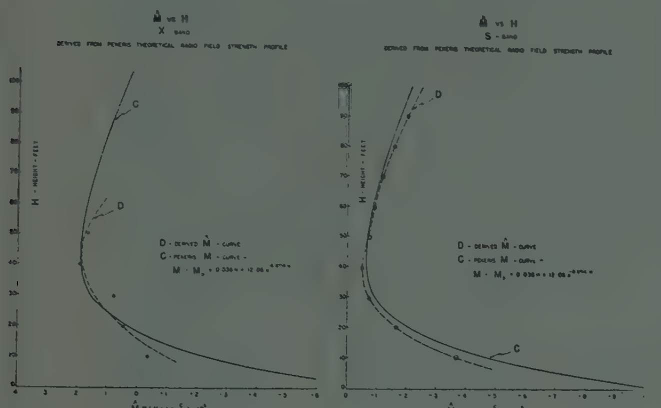
The reasonably good agreement of the derived $\widehat{M}$ curves (Fig. 1(a) and (b)) with the assumed linear M curve affirms the formal theoretical correctness of Macfarlane's method in the case of the theoretical radio-meteorological profiles of Standard Diffraction Theory.

II. The Derivation of Modified Index of Refraction Profiles from Theoretical Radio Field-Strength Data Given by Pekeris for 10-cm and 3-cm Waves Propagated in Low-Level Linear-Exponential Type Ducts

Pekeris[5] has given radio field-strength profiles for 10-cm and 3-cm waves, which he derived from a theoretical treatment assuming a linear-exponential type modified refractive index profile of the form $M(h) = Mo + 0.036h + 12.06e^{-0.0714h}$. Applying Macfarlane's method to the radio field-strength data, one should be able to derive the original assumed linear-exponential $M(h)$ profile and thus gain an insight into the behavior of Macfarlane's method under nonstandard conditions. Radio field-strength data had to be taken from the field-strength profiles given in the Pekeris paper, as no data were given there. Small transcription errors of the order of ± 0.5 decibel are inevitably incurred in such a process and as cited in the Pekeris paper, the height-gain functions were computed only approximately. The radio field-strength data were then taken from the Pekeris curves for the 3-cm and 10-cm waves for the first mode and repeatedly "smoothed."

For the 10-cm data, the average agreement between the derived $\widehat{M}$ curve and the Pekeris assumed M curve was within 0.3 M units (Fig. 2(b)). For the 3-cm case (Fig. 2(a)), only radio field-strength data between 10

and 50 feet could be obtained from the profile curve with sufficient accuracy to permit treatment. Over this height interval the average agreement was about 0.6 M units. More accurate radio field-strength data would have undoubtedly led to more consistent agreement between the derived and the assumed M curves. Macfarlane's method is then formally consistent with Pekeris' theoretical treatment of low-level linear-exponential type ducts. This is, of course, what one would expect, since both works stem from the Mode Theory of Propagation.



(a) (b)

Fig. 2—Modified refractive index profiles, derived by applying Macfarlane's method to the theoretical radio field-strength profiles given by C. L. Pekeris for a linear-exponential M curve.

III. The Derivation of Modified Index of Refraction Profiles from Experimental Overwater Radio Field-Strength Data for Radio Frequencies of 65, 170, and 520 Mc Under Essentially Standard Meteorological Conditions

Experimental overwater radio field-strength and complementary meteorological profiles have been obtained for a variety of meteorological conditions in the San Diego locality by use of a PBY-5A airplane. One-way pulse transmission was employed on 65, 170, 520 and 3,300 Mc. The radio transmitters were mounted in the airplane, and two separate sets of receivers were situated at San Diego shore locations of 100 feet and 500 feet elevation. The receiver antenna mounts were rotatable in order to permit "homing" on the airplane. The transmitting antennas were sufficiently broad to allow for minor bearing and azimuthal deviations of the airplane from the true course. The power output of each transmitter was held constant and was monitored and continuously recorded. The receivers and signal generators were run continuously during flight periods, and the receivers were calibrated before and after each flight. The calibrations were maintained within ± 0.5 db. Since ± 0.5-db error is involved in transcription of the recorded data from the recording tape to tabular form, the relative accuracy of the radio data is about ± 1.0 db. Meteorological equipment for the continuous recording of air temperature and dew point was mounted in the airplane. From these data the modified refractive index was computed by well-established procedures.[9] The radio and meteorological profiles were recorded during ascents and descents of the airplane.

An elevated temperature inversion accompanied by a lapse in water vapor pressure persists during most of the year in the locality of San Diego. This general condition is interrupted only temporarily by frontal activity. However, just after the passage of a front, meteorological conditions are often "standard" for a few hours only. On March 17, 1948, a flight was made just after the passage of a rain shower, and radio field-strength profiles were recorded on 65, 170, 520 and 3,300 Mc under essentially "standard" conditions (Fig. 3). The individual meteorological profiles exhibit several B units variation between profiles, $(B \equiv (n-1)10^6 + 0.012H)$. Evidently the atmosphere on March 17, 1948, was only approximately standard; however, it might be remarked that this is the most nearly "standard" set of data that has ever been recorded at this laboratory. The average of four meteorological soundings taken over the flight path yielded an M curve that was essentially linear between 50 and 1,200 feet and of slope, $\Delta M/\Delta H = 3.8$ M units/100 feet. This value was adopted as the control slope. The 3,300-Mc signal characteristically exhibited scintillation effects, and, as generally observed, the field strength was several decibels higher in the nonoptical region than S.D.T. predicted, which indicated the operation of some mechanism in addition to diffraction, possibly scattering.[4,5] Hence, the 3,300-Mc data was not amenable to treatment by Macfarlane's method. After subjecting the data for 65, 170, and 520 Mc to several successive "smoothings," each followed by an unsuccessful attempt to apply Macfarlane's method, it became evident that the method was entirely too sensitive to permit application directly to the experimental data, even though the data be repetitively "smoothed."[10] This sensitivity is due to the erratic behavior of the term involving the second derivative of the field strength with respect to height (d^2A/dh^2) which was computed from the "smoothed" data by the method of second differences. An alternative method of approach was to attempt to approximate the radio field-strength data, within the limits of experimental error, by an analytical curve. It was found that the standard experimental field-strength profile can, in general, be approximated satisfactorily over a pertinent portion of the height in-

[9] The relative accuracy of the experimental modified refractive index data was within ± 1.0 M unit; the relative accuracy of the measured altitudes was approximately ± 20 feet.

[10] It might be remarked that the "smoothing" of experimental data does to a certain extent render the resulting data less reliable. However, in this instance, the "smoothing" was directed at irregularities in the experimental data within the relative experimental accuracy of that data, i.e., irregularities of less than ± 1 decibel. For the present considerations, significant changes in the experimental data due to propagation anomalies are of the order of several decibels. Therefore, it is believed that no perturbations in the experimental data of significant magnitude were eliminated by the process of "smoothing." The same remarks apply to the approximation of the experimental data by analytical curves.

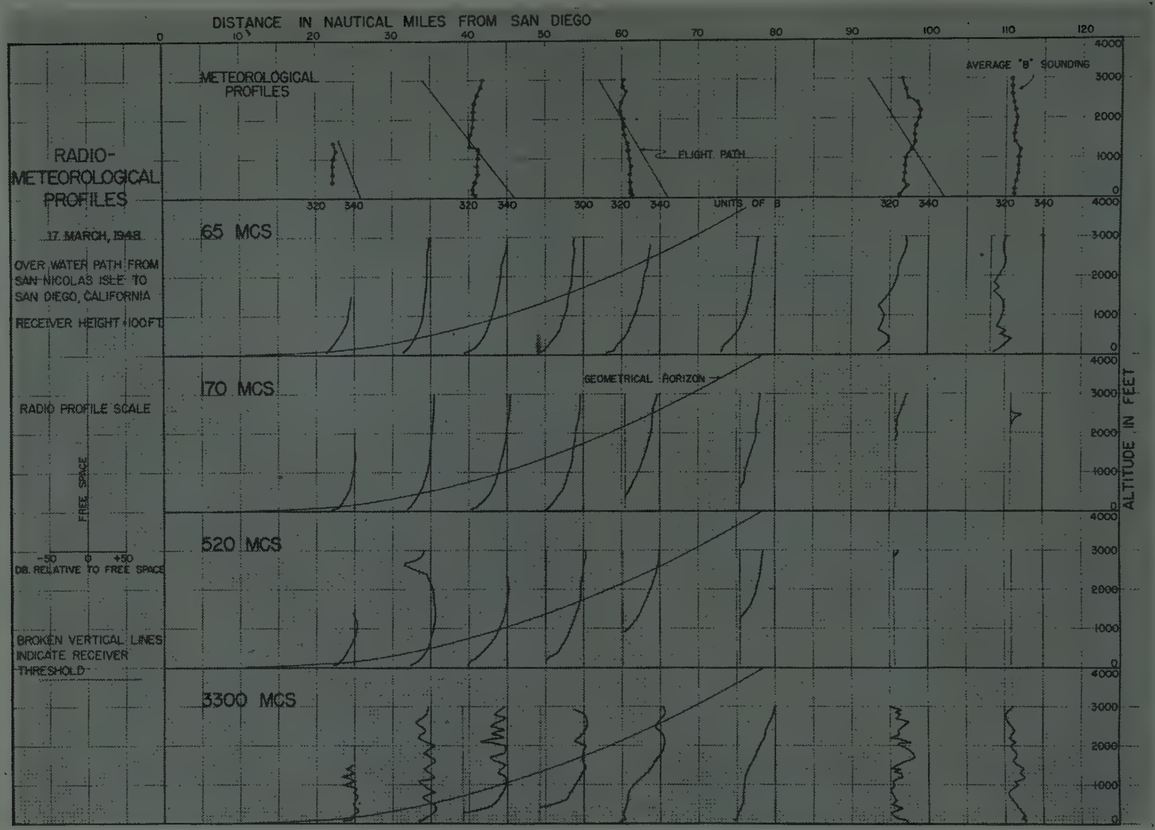
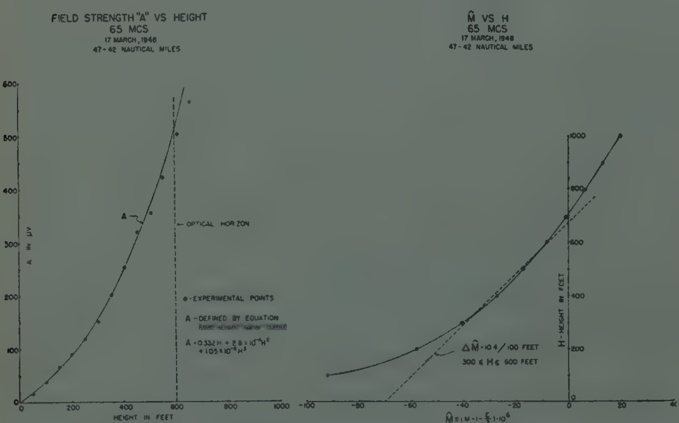


Fig. 3—Radio-meteorological profiles recorded on an overwater transmission path on March 17, 1948, under essentially standard lapse rate conditions. $(B \equiv n-1)10^6 + 0.012H$.

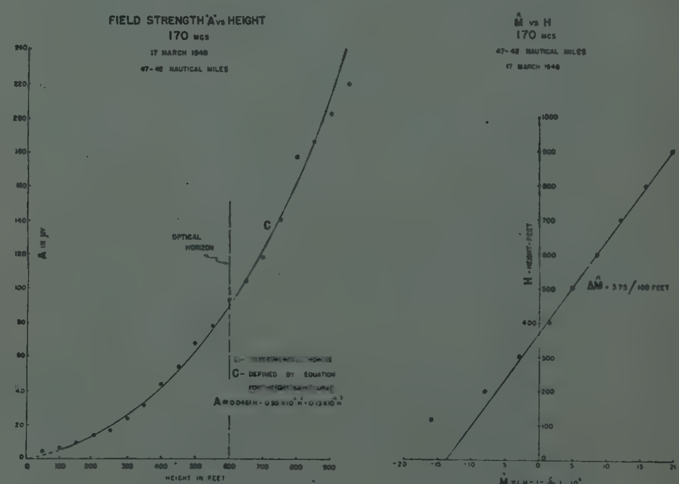
terval by a curve of the second degree, $A = a + bH + cH^2$. Somewhat better agreement may often be obtained by using a linear exponential type curve of the form, $A = a + bH + ce^{\pm nH}$ or a power function of the form, $A = a + bH + cH^{\pm n}$, where n may be nonintegral. The expression d^2A/dh^2 computed from such an approximating function is then well behaved and found to lead to consistent $\widehat{M}$ data. The results of this approach applied to 65, 170, and 520 Mc profiles are shown in Figs. 4(a), 5(a), and 6(a), together with the derived $\widehat{M}$ curves, Figs. 4(b), 5(b), and 6(b). The radio profiles used in the computation were taken at a distance of 47 to 42 nautical miles from the receiver location. The attenuation constant, $kb/2$, was computed using values of field-strength taken from the nonoptical portion of neighboring profiles. The optical horizon was at approximately 600 feet, so only data up to 600 feet were considered for treatment.

The agreement between the values of $\Delta\widehat{M}/\Delta H$ for the 170- and 520-Mc cases was within 12 per cent. The average value of $\Delta\widehat{M}/\Delta H$ for 170 and 520 Mc was 3.53 /100 feet and agreed with the average measured control slope, $\Delta\widehat{M}/\Delta H = 3.8/100$ feet, within 7 per cent. The



(a) (b)

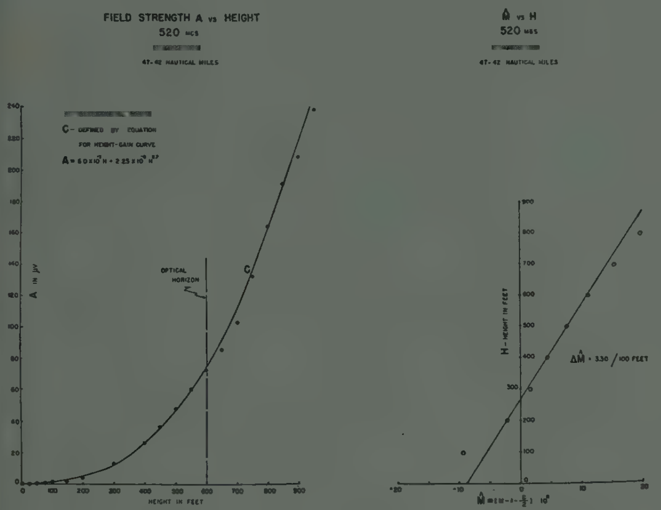
Fig. 4—(a) Plot of 65-Mc experimental radio field-strength data and an approximating curve. (b) Derived $\widehat{M}$ curve.



(a) (b)

Fig. 5—(a) Plot of 170-Mc experimental radio field-strength data and an approximating curve. (b) Derived $\widehat{M}$ curve.

derived values of $\widehat{M}$ for 65 Mc did not form a straight line. A "best" line through the nonoptical region possessed a slope of $\Delta M/\Delta H = 10.4/100$ feet or virtually no agreement with the average observed value. Thus, because of an inherent frequency effect (see Discussion), the method applied to experimental data fails to provide a dependable answer for frequencies less than about 100 Mc. However, because of the presence in general of scintillations in the radio field-strength on frequencies above about 1,000 Mc, it appears that the practical range of application of the method is limited to frequencies between 100 and 1,000 Mc.



(a) (b)

Fig. 6—(a) Plot of 520-Mc experimental radio field-strength data and an approximating curve. (b) Derived $\widehat{M}$ curve.

It is not apparent, however, that Macfarlane's method provides information in the case of standard propagation which could not more readily be obtained from Standard Diffraction Theory.[11]

IV. Derivation of Modified Index of Refraction Profiles from Experimental Overwater Radio Field-Strength Profiles for Nonstandard Conditions

A set of nonstandard radio field-strength data (Fig. 7), recorded on April 8, 1948 on an overwater path by use of the PBY airplane, was selected for treatment by Macfarlane's method. Over the entire transmission path of 280 nautical miles, an elevated layer was present which had an average base height of 600 feet, an average thickness of 400 feet, and an average M deficit of 42 M units. However, horizontal atmospheric stratification only approximately prevailed over the transmission path, though in this instance, a better approximation existed than is usually the case. As is generally observed, the base of the inversion lowered markedly at some 75 nautical miles off shore. In this case the base dropped from 700 feet at 93 nautical miles from San Diego to 400 feet at 56 nautical miles from San Diego. However, in this instance, the inversion base from 93 nautical miles to 260 nautical miles remained substantially constant at 700 feet. In general, the refractive index discontinuity takes the form of a warped surface, the gradients varying from point to point, and rarely does the lower discontinuity surface remain as uniform over such an extensive distance as was measured on

[11] For example, by inverse interpolation of one of the several sets of published nomograms on S.D.T.

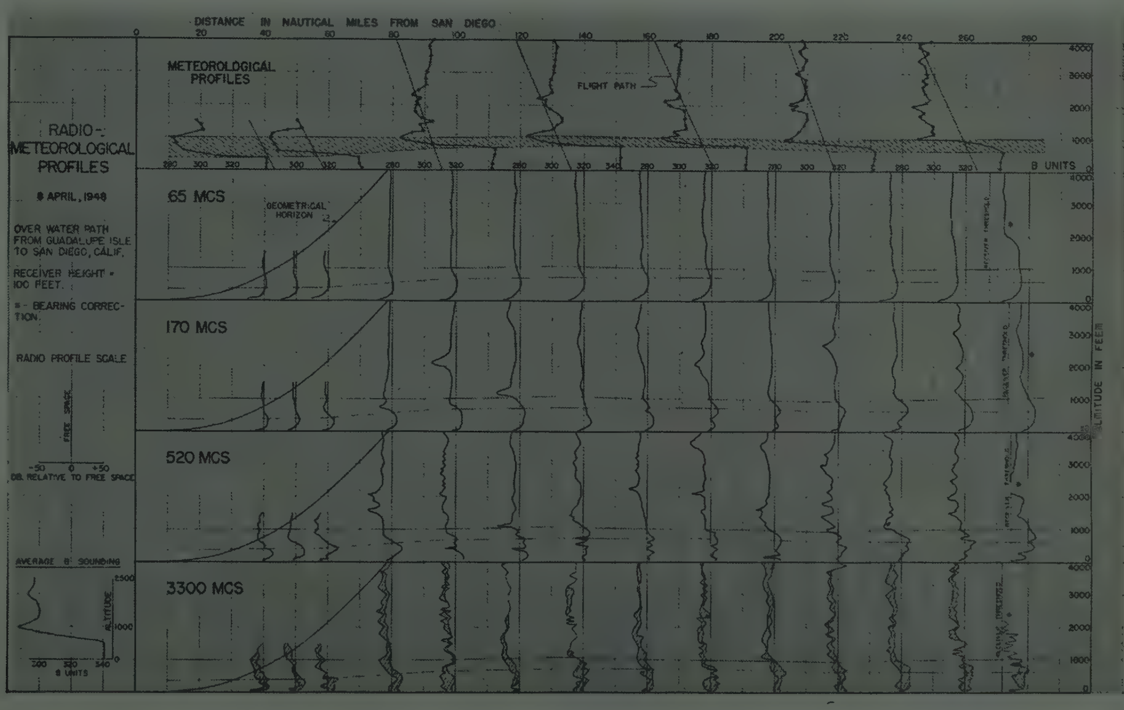
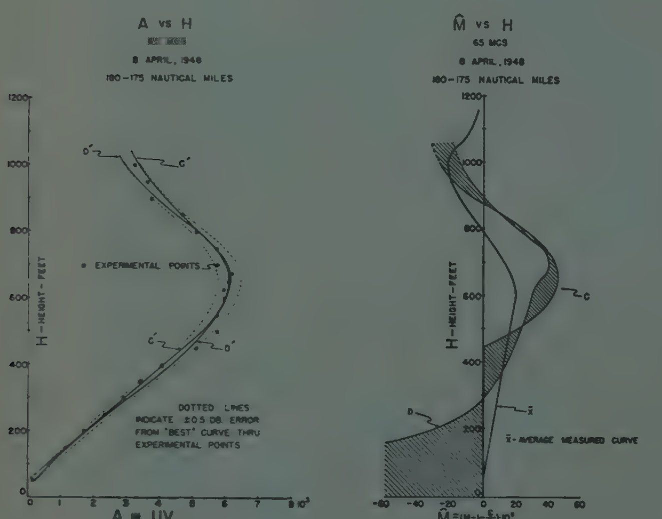


Fig. 7—Nonstandard radio-meteorological profiles recorded on an overwater transmission path on April 8, 1948. $B = (n-1)10^6 + 0.012H$.

April 8, 1948. Reference to Fig. 7 suggests that more than one mode was operative, even at distances beyond 200 nautical miles, on the frequencies of 170, 520 and 3,300 Mc. As usual, scintillations of several decibels' magnitude were present on the 3,300-Mc signal, and the profiles shown are envelopes of the maximum and minimum field strength. However, since the shape of the height-gain profile of the 65-Mc signal was approximately independent of distance to about 200 nautical miles, it appeared that essentially single mode propagation prevailed on that frequency.

A 65-Mc radio profile recorded from 180 to 170 nautical miles (Fig. 7), was considered for treatment by Macfarlane's method. An average value of the attenuation constant $kb/2$ was obtained by considering several of the profiles recorded between 100 and 180 nautical miles. Previous experience indicated that the radio data could not be used directly but would have to be approximated within the limits of experimental error by one or more analytical curves, joined together appropriately so that the functions and their second derivatives were continuous at the point of juncture. It was decided to make an exploratory analysis of the first 1,000 feet of the radio profile. If preliminary work with the first 1,000 feet of the radio profile yielded successful results, more junctures could be effected, and approximating curves found to carry the analysis to any desired height. Further work above 1,000 feet would not, of course, effect the derived refractive index profile below 1,000 feet. The experimental 65-Mc radio profile data were approximated, within the limits of experimental error, by a straight line from 100 to 450 feet, joined at 450 feet to a sinusoid from 450 to 1,000 feet, curve C' of Fig. 8(a). The resulting derived $\widehat{M}$ curve is curve C (Fig. 8(b)), and is only in rather poor qualitative agreement with curve $\overline{X}$, which, is the average measured M curve. It



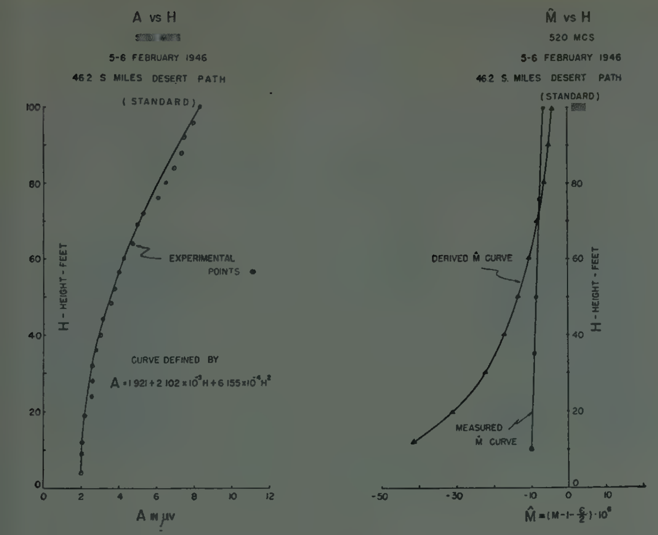
(a) (b)

Fig. 8—(a) Plot of 65-Mc experimental radio field-strength profile data and two approximating curves C' and D'. (b) The derived modified refractive index profiles, C and D, and the average measured profile $\overline{X}$.

appeared that the 65-Mc experimental radio field-strength data under consideration might be approximated equally well, within the limits of experimental error, by two-third degree curves appropriately joined together. The resulting approximating curve is D' (Fig. 8(a)). The resulting derived $\widehat{M}$ curve is curve D (Fig. 8(b)), which again is in rather poor qualitative agreement with the average measured M curve, $\overline{X}$. The dotted lines (Fig. 8(a)) indicate ± 0.5 db error from the "best" curve through the experimental points. By choosing a variety of other approximating curves for the radio data, *all within the limits of experimental error*, a variety of derived $\widehat{M}$ curves lying within or near the cross-hatched region (Fig. 8(b)) would result; which one, if any, was the representative one, it would be impossible to decide. Treatment of other 65-Mc radio profiles resulted in typically unsatisfactory $\widehat{M}$ curves. It might be remarked that of the many sets of nonstandard radio-meteorological data recorded at this Laboratory during the past two years, the 65-Mc radio data recorded on April 8, 1948, offered the most promise of being amenable to treatment by Macfarlane's method. Under the more typical nonstandard meteorological conditions prevalent in this locality, both the height and thickness of the elevated layer change markedly with distance. Under these typical conditions the shape and character of all the radio profiles, including the 65-Mc, are a function of distance.

V. The Theoretical Deduction of Modified Index of Refraction Profiles from Experimental Overland Radio Field-Strength Profiles for Standard and Nonstandard Conditions

Experimental radio field-strength profiles for 170, 520, 1,000, 3,300, 10,000 and 30,000 Mc have been recorded under standard and nonstandard meteorological conditions over a 46.2-statute mile overland path in the Arizona desert.[6] In general, scintillations several decibels in magnitude were present on 1,000-, 3,300-, 10,000- and 30,000-Mc profiles. Minima in the radio height gain profiles, often 20 decibels or more in magnitude,[6] were present on 3,300, 10,000 and 30,000 Mc and to a lesser extent and degree of magnitude on 170, 520, and 1,000 Mc, under both standard and nonstandard meteorological conditions. The cause of this effect is not understood. However, more than one mode of propagation, nonhomogeneous atmospheric stratification, and possibly diffraction effects due to small prominences in the otherwise smooth terrain, probably contribute to the effect. In general, the shapes of the radio height-gain profiles of all the frequencies recorded are much more complicated than would be expected on the basis of an idealized theory assuming single mode propagation and a simple horizontally stratified atmosphere. However, the lower segments of certain of the 520-Mc radio profiles appeared to be smooth enough to admit treatment by Macfarlane's method. The lower portion of a 520-Mc profile recorded on February 5, 1948, with a transmitter

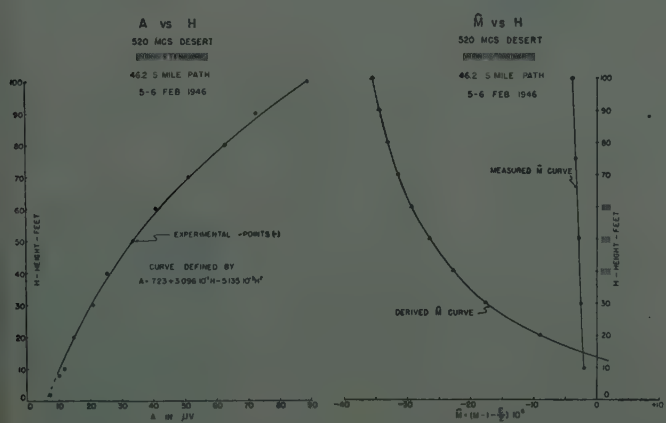


(a) (b)

Fig. 9—(a) Plot of experimental 520-Mc radio field-strength data and an approximating curve. (b) Derived modified refractive index profile and the average measured profile.

height of 100 feet, under essentially standard meteorological conditions, is shown in Fig. 9(a), together with an approximating curve and the defining analytical expression, which approximated the measured data within the limits of experimental error. The resulting theoretically derived $\widehat{M}$ curve together with the prevailing average measured M curve (linear: 3.7 M units/100 feet) are shown in Fig. 9(b). The derived $\widehat{M}$ curve was not linear, and hence no representative slope could be assigned to it.

Fig. 10(a) shows the lower segment of a 520-Mc radio field-strength profile recorded on February 6, 1946, with a transmitter height of 100 feet, under the presence of a ground-based radiation inversion. An approximating curve and the defining analytical expression are also shown in Fig. 10(b). The resulting theoretically derived $\widehat{M}$ curve is shown in Fig. 10(b), together with the prevailing average measured M curve. The derived $\widehat{M}$ curve had entirely too large a negative slope and could



(a) (b)

Fig. 10—(a) Plot of nonstandard experimental 520-Mc radio field-strength data and an approximating curve. (b) Derived modified refractive index profile and the average measured profile.

not be considered representative of the prevailing refractive index distribution. Treatment of several other selected 520-Mc radio profiles and one case for 1,000 Mc have all yielded typically unsatisfactory results. Thus, attempts to describe prevailing refractive index distributions over the Arizona desert propagation path by Macfarlane's method have been unsuccessful, even though radio field-strength profile data were selected which were relatively smooth and appeared to be free from the generally present perturbations alluded to previously.

Discussion and Conclusions

The application of Macfarlane's method to the theoretical field-strength profiles of Standard Diffraction Theory, where there is an established relationship between the field-strength and the meteorological profiles, revealed that in order to theoretically derive satisfactory refractive index profiles, the accuracy of the radio field-strength data must be extremely high (of the order of ± 0.1 db) in order to apply the method directly to the field-strength data. For frequencies above 1,000 Mc, somewhat less accuracy would be acceptable, while for frequencies less than 1,000 Mc, greater accuracy would be necessary. The reason for this sensitivity is the relationship between the two terms in the expression for $\widehat{M}$ (6). Because of the second derivative term, the expression $1/k^2A\ d^2A/dh^2$ is extremely sensitive to slight inaccuracies in the field-strength data; further, since the magnitude of this expression is inversely proportional to frequency squared, the effect of small errors in the field-strength data will have more adverse effects at lower frequencies ($f<1,000$ Mc) than at higher frequencies ($f>1,000$ Mc). However, the magnitude of the expression

$$\frac{k^2b^2}{A^4}\left\{\int_0^H A^2dh\right\}^2$$

is directly proportional to the frequency squared, and further, the integral expression

$$\left(\int_0^H A^2dh\right)$$

is relatively insensitive to small errors in the field-strength data. The expression

$$\frac{1}{k^2A}\,d^2A/dh^2$$

which is the more sensitive, is dominant over the expression

$$\frac{k^2b^2}{A^4}\left[\int_0^H A^2dh\right]^2$$

in those segments of the radio height-gain curve where the curvature is changing at an appreciable rate, e.g., the lower segment of the radio profile, near turning points and inflection points of nonstandard height-gain

profiles. However, it is in the neighborhood of just such points that, in general, the most significant changes in the complementary refractive index profile occur. Unfortunately, it is in these regions that the derived $\widehat{M}$ curve can least accurately be determined. It might be remarked that similar difficulties have been experienced in attempts to apply the phase-integral method to the determination of the characteristic values and normalization factors of the wave equation for various assumed refractive index profiles.[12] The gaps between the regions of validity of the phase-integral approximation turn out to be extremely wide and to cover just the more critical and interesting ranges of slope and duct height. Macfarlane's method appears to be subject to the same sort of difficulty, which is not surprising since it is essentially a phase-integral method of solution of the wave equation for the expression $(a-M^2)$.

The accuracy of experimental radio field-strength data (± 1.0 decibel) is insufficient to permit the application of Macfarlane's method directly to the data, even after repeated "smoothing." However, in the case of field-strength data recorded under standard meteorological conditions, it appeared to be possible to approximate, within the limits of experimental error, the lower segment of the experimental field-strength profile by a second degree curve, and then carry out the computations using the approximate curve. In some cases, a power function of the height or an exponential function appeared to offer better approximations. Macfarlane's method then applied, using the values of the radio field-strength defined by the approximating analytical curve, yielded relatively consistent modified refractive index data for the frequencies of 170 and 520 Mc (Fig. 5(b) and Fig. 6(b)). However, reliable results could not be expected for frequencies less than about 100 Mc. Because of the presence, in general, of scintillations in the radio field strength for frequencies above 1,000 Mc, the method would not be applicable. For standard (linear) lapse rate conditions, it was not apparent, however, that Macfarlane's method would provide information, within the limited frequency range 100 to 1,000 Mc, that could not more readily be obtained from Standard Diffraction Theory.

Because of their more complicated shape, it is much more difficult to approximate nonstandard radio field-strength profiles by analytical curves. In general, two or more separate functions must be used over different portions of the profile. The functions and their second derivatives at the points of juncture must be made continuous, and of course the functions must approximate the experimental data within the limits of experimental error. The position of inflection points in the radio field-strength profile, which are extremely important in the determination of the $\widehat{M}$ curve, are difficult to determine because of the relative inaccuracy of experimental data.

The assumptions of horizontal stratification and single mode propagation, if not strictly fulfilled, render the solution

$$W = \alpha U(h)e^{-ikx\cos\theta} \tag{7}$$

only an approximation. Essentially it is the first term of a Fourier expansion for the wave function W. In all practical cases, then, (7) is an approximate solution; hence, the theoretical amplitudes required to satisfy relation (7), are only approximations to the experimentally measured amplitudes. How good an approximation they are is dependent upon how nearly the assumptions of horizontal stratification and single mode propagation are fulfilled. However, even in instances where the theoretical amplitudes *are very satisfactory approximations* to the measured amplitudes, the second derivatives of the measured amplitudes, by the mathematical nature of the process of differentiation[13] will, in general, be in *unsatisfactory approximate agreement* with the second derivatives of the theoretical amplitudes. Thus, in the realm of experience, particularly for nonstandard meteorological conditions, one would, in general expect the refractive index profile derived by Macfarlane's method to be an unsatisfactory approximation to the prevailing average measured profile. Attempts to apply the method, particularly in the case of nonstandard experimental data, have indeed shown this to be the case. It might be emphasized that this difficulty would still exist, even though a higher degree of experimental accuracy in the measurement of the experimental radio field-strength amplitudes were attainable, since perturbations of the radio field would, in general, be present, due to the fact that the atmosphere is only approximately horizontally stratified and that contributions from other modes, higher than the first, may often be present.

In any given instance it might be difficult, perhaps impossible to distinguish those perturbations of the observed radio field-strength profile due to the presence of more than one mode of propagation, from those due to the effects of nonhorizontal atmospheric stratification. In any event, such perturbations are always present to a greater or lesser extent in any practical experiment. The solution (7) then, only approximately represents the observed radio field.

With the possible exception of a limited frequency range under linear lapse rate conditions, it appears that Macfarlane's method is unreliable and inadequate for the description of experimentally observed propagation phenomena.

Acknowledgments

The author wishes to express appreciation to his colleagues, R. A. Baker, M. D. Rocco, and C. P. Hubbard for assistance in performing the laborious computational work, and in particular, to J. B. Smyth for helpful suggestions and continued encouragement during the progress of the work.

[12] W. H. Furry, "Theoretical Treatment of Non-Standard Propagation in the Diffraction Zone," Third Conference on Propagation Notes, November, 1944: Committee on Propagation, NDRC of the OSRD.

[13] The author is indebted to Dr. I. S. Sokolnikoff of the Mathematics Department, University of California at Los Angeles, for pointing out that the derivatives of an approximation of a function are, in general, very unsatisfactory approximations of the derivatives of the function itself.

Contributors to the Proceedings of the I.R.E.

Donald F. Aldrich

Donald F. Aldrich (A'46) was born on December 12, 1923, in Milwaukee, Wis. He received the B.Sc. degree in electrical engineering from the University of Michigan in June, 1945. Immediately after graduation he joined the Westinghouse Electric Corporation as an engineer in the industrial electronic control department. In July, 1948, Mr. Aldrich transferred to the Westinghouse Research Laboratories, where he has since been principally engaged in studies of magnetic amplifiers. He continued study at the Carnegie Institute of Technology during this period, and was awarded the M.S. degree in electrical engineering in June, 1948.

❖

John H. Bollman

John H. Bollman (M'42–SM'43) was born in Hoboken, N. J., in 1906. He attended the Polytechnic Institute of Brooklyn, receiving the B.E.E. degree in 1939. In 1922 he joined the engineering department of the Western Electric Company, which became the Bell Telephone Laboratories in 1925. He is now a member of the technical staff of the Bell Telephone Laboratories, where he has been engaged in the development of carrier telephone repeaters and, more recently, of semiconductor devices such as thermistors.

Mr. Bollman is a member of the American Physical Society, American Association for the Advancement of Science, and Eta Kappa Nu.

❖

K. Bullington

Kenneth Bullington (A'45) was born at Guthrie, Okla., on January 11, 1913. He received the B.S. degree in electrical engineering in 1936 from the University of New Mexico, and the M.S. degree in 1937 from the Massachusetts Institute of Technology. Since his graduation he has been a member of the technical staff of the Bell Telephone Laboratories, where he is engaged in wire and radio transmission problems.

E. Eugene Ecklund

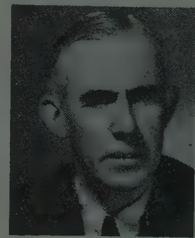
E. Eugene Ecklund was born in Akeley, Minn., on January 16, 1921. He received the B.E.E. degree from the University of Minnesota in 1942. Then, on active duty with the U. S. Navy until February, 1946, attaining the rank of Lieutenant, USNR, he served as assistant head of the test equipment section, Bureau of Aeronautics Radio and Electrical Engineering group, and as technical intelligence officer with the Naval Air Attaché's Office, London, and the U. S. Naval Technical Mission in Europe, conducting field liaison work with USSTAF Air Technical Intelligence Unit and combined Intelligence Objectives Sub-committee (CIOS). Upon release from active duty, he became manager of the National Service Department of Allen B. DuMont Laboratories, Inc., which he left in October, 1946, to organize Bergen-Passaic Electronics, Inc., serving in his present capacity as president.

❖

James W. Green

James W. Green (A'47) was born in Colorado Springs, Colo., on March 11, 1912. He received the A.B. degree in physics from Colorado College in 1936, and the M.A. degree in physics in 1938 from Dartmouth College, Hanover, N. H. During the period 1938 to 1940 he continued graduate work in physics at the California Institute of Technology, Pasadina, Calif., on a fellowship grant from the Cramer Foundation. In 1941, Mr. Green joined the research and development staff of RCA Radiation, Inc., Harrison, N. J., where he assisted with the design and development of megawatt pulsed transmitting tubes.

From 1942 to 1945 Mr. Green was employed at the Naval Research Laboratory, Washington, D. C., where he assisted with airborne radar, radar counter-measures, and high-frequency direction-finder developments. Since 1946, he has been engaged in wave-propagation research work at the U. S. Navy Electronics Laboratory, San Diego, Calif., where he is a physicist in the Atmospheric Studies Branch of the Research Department. He is a member of the Society for Research on Meteorites and the American Association for the Advancement of Science.



Hans E. Hollmann

Hans Erich Hollmann (A'48) was born on November 4, 1899, at Solingen, Germany. He majored in physics at the Technical University of Darmstadt in 1920, and received the degree of doctor in 1928. His thesis was concerned with the generation of microwaves as far down as 20 cm with an energy of about 1 watt. He continued his investigations on very-high frequency as a scholar of the "Notgemeinschaft der Deutschen Wissenschaft" (Emergency Society of German Science), transferring in 1930 to the Heinrich-Hertz Institut für Schwingungsforschung in Berlin. During 1934–1936 he wrote the first original encyclopedia concerning vhf and microwaves, including a preliminary description of the phenomenon later known as *radar*. Since then, he has been occupied with extensive studies and investigations relative to microwaves, vhf diathermy, and electrocardiography (his so-called "Triography").

Numerous scientific papers were published in Germany, England, and the United States from his privately owned Laboratory for High-Frequency Techniques and Electromedicine. At the same time, Dr. Hollmann was in charge of development and research as consulting engineer for the "Telefunken," "Siemens," and other companies. During this period, he originated important inventions concerning radar and beam-riding. In 1941 he became scientific director of the Research Society for Radio and Movies. After the war, Dr. Hollmann became professor of high frequency and electromedicine at the Friedrich Schiller University in Jena, Germany. In 1947 he left Germany for the United States.

❖

Luther W. Hussey

Luther W. Hussey (A'41–SM'46) was born in Milford, Mass., on March 11, 1901. He received the A.B. degree from Dartmouth College in 1923, and the M.A. from Harvard University in 1924. After six years' teaching experience at Union College, in Schenectady, N. Y., he joined the technical staff of the Bell Telephone Laboratories, where he has been engaged in research on nonlinear resistive and reactive materials, and their circuit applications.

Heinz E. Kallmann (A'38–M'41–'SM'43) was born on March 10, 1904, in Berlin, Germany, He received the Ph.D. degree in physics from the University in Goettingen in 1929. From 1929 to 1934, Dr. Kallmann was a research engineer in the laboratories of the C. Lorenz A. G. in Berlin, concerned with the development of rf test equipment, and of ultra-short-wave receivers for communications and for wide-band television. He was also in charge of television development.

H. E. Kallmann

From 1934 to 1939, Dr. Kallmann was a television engineer wtih Electric and Musical Industries, Ltd., Hayes, England, doing research and advanced circuit development. Since 1939, he has been a consulting engineer in New York, N. Y., except for the period from 1943 to 1945, when he joined the Radiation Laboratory of the Massachusetts Institute of Technology, where he had charge of the circuit section of the test equipment group, and was a member of the fundamental microwave research group.

Dr. Kallmann is a member of the American Physical Society; an associate member of the Society of Motion Picture Engineers, and a member of its committee on television projection practice; and a member of the RMA committee on uhf television systems and some of its subcommittees.

❖

John J. Karakash was born in Constantinople, Turkey, in 1914. He attended Robert College, and in 1936 he received the Angier Duke Scholarship at Duke University, graduating with a B.S.E.E. in 1937. He became a graduate fellow at the Moore School of Electrical Engineering, University of Pennsylvania, where he obtained the master's degree in June, 1938. Prior to his present association with Lehigh University, where he serves as an assistant professor of electrical engineering, Mr. Karakash served on the teaching staff and as a research engineer at the Moore School of Electrical Engineering. He is the author of a text titled "Transmission Lines and Filter Networks," now in the process of publication (The Macmillan Company). He is the co-winner of the first (1949) Nobel Robinson $1000 Award given at Lehigh University to a young member of the faculty for outstanding service.

John J. Karakash

❖

For a photograph and biography of A. G. Clavier, see page 409 of the April, 1949, issue of the Proceedings of the I.R.E.

J. G. Kreer (M'40–SM'43) was born at Chicago, Ill., on March 25, 1903. He received the B.S. degree in electrical engineering from the University of Illinois in 1925, and the M.A. degree from Columbia University in 1928. Since 1925 he has been a member of the technical staff of the Bell Telephone Laboratories, where he has worked on problems connected with nonlinear circuits and carrier telephone systems development.

J. G. Kreer

Mr. Kreer has served on the IRE Papers Review Committee, the Membership Committee, the Modulation Systems Committee, and the Annual Review Committee.

❖

A. H. LaGrone (M'48) was born in Panola County, Texas, on September 25, 1912. He received the degree of B.S. in electrical engineering from the University of Texas in 1938. After four years as distribution engineer with the San Antonio Public Service Co., San Antonio, Texas, he was commissioned in the U. S. Naval Reserve and was ordered to active duty with the Navy in June, 1942. While in the Navy, Mr. LaGrone was Instructor in Radar at the Massachusetts Institute of Technology and later Radar Officer aboard the U.S.S. *Gillette*, D. E. 681, in the Atlantic.

A. H. LaGrone

At the conclusion of the war, Mr. LaGrone, then a Lieutenant Commander, was ordered to inactive duty and accepted the position of radio engineer with the Electrical Engineering Research Laboratory, the University of Texas. Mr. LaGrone also attended the University of Texas and was awarded the degree of M.S. in electrical engineering in 1948.

Mr. LaGrone is a member of Eta Kappa Nu and Tau Beta Pi.

❖

Matthew T. Lebenbaum (A'42–M'46–SM'46) was born in Portland, Ore., on November 29, 1917. He received the B.A. degree from Stanford University in 1938, after which he accepted a post as research and teaching assistant at the Massachusetts Institute of Technology. He received the M.S. degree in 1945. After one year with the American Gas and Electric Service Corporation, he joined the Radio Research Laboratory of Harvard University in 1942 as a research associate in the receiver group, and participated in the countermeasures receiver development program at Harvard and in England with the TRE and ABL-15. In late 1945 he joined the Airborne Instruments Laboratory as assistant supervising engineer of the receiver section.

M. T. Lebenbaum

Mr. Lebenbaum is a member of Sigma Xi, Tau Beta Pi, and Phi Beta Kappa.

❖

Douglas E. Mode (SM'46) was born on April 4, 1911, in Brandon, Manitoba, Canada. He was graduated from the University of Pennsylvania, Moore School of Electrical Engineering, with the degrees of B.S. in 1935, M.S. in 1937, and Ph.D. in 1947. He joined the engineering staff of the General Electric Co., in 1936, where he was engaged in the design and development of magnetic and electronic relaying devices.

Douglas E. Mode

In 1940 Dr. Mode became an instructor in electrical engineering at Lehigh University and was made an assistant professor in 1941, in which capacity he served until early 1944. On leave from Lehigh University, he joined the Columbia University Underwater Sound War Research Group at New London, Conn. later transferring to the Radiation Laboratory at MIT.

Returning from war work, Dr. Mode again took up teaching duties at Lehigh University, where he is now an associate professor of electrical engineering. He is a member of Eta Kappa Nu, Tau Beta Pi, and Sigma Xi.

❖

John Ruze (S'39–A'40–M'46) was born in New York, N. Y., on May 24, 1916. He received the B S. degree in electrical engineering from the College of the City of New York in February, 1938, and the M.S. degree from Columbia University in June, 1940. He became associated with the Signal Corps Engineering Laboratories upon graduation, and from 1942 to 1946 he headed the antenna design section at the Evans Signal Laboratory. In June, 1946, Mr. Ruze joined the Air Force Cambridge Electronic Research Laboratory, where he served as assistant chief of the Antenna Laboratory until June, 1948. At the present time he is associated with the Radar Laboratory of that organization.

John Ruze

Mr. Ruze is a member of the American Physical Society, Epsilon Chi, Tau Beta Pi, and Eta Kappa Nu.

John M. Shaull was born in Hagerstown, Md., on August 31, 1910. He attended public schools in Charles Town, W. Va., and was occupied as radio and electrical serviceman there until 1935. He has held an amateur radio license since 1930, and commercial radiotelephone since 1937. In 1935 he came to Washington, D. C. and worked at several Government departments, during which time he graduated from Capitol Radio Engineering Institute, and attended George Washington University. In 1938 he was employed in the engineering department of Bendix Radio Corporation, Baltimore, Md., calibrating and adjusting precise frequency measurement equipment.

John M. Shaull

In 1939 Mr. Shaull joined the staff of the National Bureau of Standards, serving as operator at the Bureau's standard frequency-radio station WWV. From 1941 to 1943 he designed and helped to construct and install the frequency and time-interval control equipment for the new WWV station. The following year he designed and supervised the construction of the Bureau's microwave frequency standard completing the project in 1946. Since then he has been responsible for monitoring the accuracy of the WWV standard frequency and time transmissions and improving the standards and methods associated with and constituting the primary standard of frequency.

❖

Harold W. Smith (A'48) was born in Brookfield, Mo., on February 8, 1923. He entered the University of Texas in 1941, receiving the B.S. degree in electrical engineering in 1944, and the M.S. degree in 1949. From 1944 to 1946 he served as an Airborne Radar Officer in the U. S. Navy. Since 1946 he has been an instructor in electrical engineering and a staff member of the Electrical Engineering Research Laboratory at the University of Texas.

H. W. Smith

Mr. Smith is a member of Tau Beta Pi and Eta Kappa Nu.

R. L. Smith-Rose (SM'44–F'45) was born in London, England, in 1894. He received his scientific and technical education at the Imperial College of Science, London University, where he obtained the B.S. degree with first-class honors in physics in 1914, and the diploma of the Imperial College in electrical engineering in 1915. These courses were followed by four year's practical experience with Messrs. Siemens Bros. and Company Limited.

R. L. Smith-Rose

Dr. Smith-Rose joined the scientific staff of the National Physical Laboratory at Teddington, Middlesex, as a member of the Electricity Division in 1919, and later formed the nucleus of the Radio Division. He has been associated with the work of the Radio Research Board of the Department of Scientific and Industrial Research, London, since the formation of the Board in 1920, and has been responsible for conducting extensive investigations in radio direction finding, the electrical properties of soil and sea water, and the propagation of radio waves over the ground and through the lower atmosphere. In the course of this research work, Dr. Smith-Rose received the Ph.D. and D.S. degrees at London University.

In 1939, Dr. Smith-Rose was appointed superintendent of the Radio Division of the National Physical Laboratory, and in 1948, he became Director of Radio Research in the department of Scientific and Industrial Research, England. He is a member of the British National Committee for Scientific Radiotelegraphy, and has attended various international conferences.

Dr. Smith-Rose is a Member of the Institution of Electrical Engineers, London, of which he was Chairman of the Radio Section in 1943, and is the recipient of five Premiums for original papers read before the institution. He received the IRE Fellow award in 1945 "in recognition of his pioneer work in the field of direction finding and radio propagation allied to his leadership of an outstanding radio research group," and he served as Vice-President of The Institute of Radio Engineers in 1948. Dr. Smith-Rose was awarded the U. S. Medal of Freedom with Silver Palm in August, 1947, for exceptionally meritorious service in scientific research and development.

❖

For a photograph and biography of D. D. King, see page 779 of the July, 1949, issue of the Proceedings of the I.R.E.

A. W. Straiton (M'47) was born in Tarrant County, Texas, on August 27, 1907. He received the B.S. degree in electrical engineering in 1929, the M.A., in 1931, and the Ph.D. in 1939, all from The University of Texas.

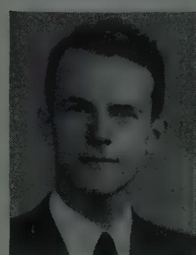
A. W. Straiton

Dr. Straiton spent one year at Bell Telephone Laboratories, after which he taught at Texas College of Arts and Industries as assistant professor, associate professor, and professor of electrical engineering, successively. From 1941 to 1943, he was head of the Department of Engineering, Institutional Representative of E.S.M.W.T., and director of the Pre-Radar Training courses. Since 1943, he has been associate professor of electrical engineering at The University of Texas. He was recently made director of the Electrical Engineering Research Laboratory.

Dr. Straiton is a member of Sigma Xi, Tau Beta Pi, Eta Kappa Nu, the American Institute of Electrical Engineers, and the American Society for Engineering Education.

❖

James B. Woodford, Jr., (S'46) was born in Claremont, Calif., in 1928. Since 1945 he has been a student at the Carnegie Institute of Technology, where he received the B.S. degree in 1948 and the M.S. degree in 1949, both in electrical engineering.



J. B. Woodford, Jr.

Mr. Woodford was for three years the recipient of a George Westinghouse Scholarship, and was employed by the Westinghouse Electric Corporation during the summer of 1947. He was awarded the Charles LeGeyt Fortescue Fellowship (AIEE) for the years 1948–1949 and 1949–1950. He is a member of Sigma Xi, Eta Kappa Nu, and Tau Beta Pi.

❖

For a photograph and biography of Everard M. Williams, see page 902 of the August, 1949, issue of the Proceedings of the I.R.E.

Institute News and Radio Notes

1950 IRE National Convention Slated in New York March 6–9

"Behind the Scenes in Radio-Electronics" is the theme of the 1950 IRE National Convention scheduled for March 6 to 9 at the Hotel Commodore and the Grand Central Palace in New York City. Manufacturers' displays of outstanding developments and products of electronic research will be shown at the Grand Central Palace under the general exhibit title, "Spot-lighting the New."

A varied and informative technical program is planned, comprised of numerous sessions devoted to contributed papers. In addition, seven especially planned symposia will be presented to Institute members. Five symposia will be sponsored by the Professional Groups in the fields of their specialties. Professional Groups are felt to be in a singularly favorable position to undertake the organization of symposia, since the primary concern of each is a particular subject with which members are intentionally concerned.

The symposia and their sponsors have scheduled tentative programs as follows: Network Synthesis in the Time Domain, Circuit Groups; Nuclear Science and the Radio Engineer, Nuclear Science Group; Industrial Design, Broadcast and Television Receiver Groups; Engineering for Quality in Television, Quality Control Group; Noise and Distortion in Sound Recording Transmission Systems, Audio Group; Basic Circuit Elements, Committee on Instruments and Measurements; Television, Special Committee.

Social events slated in connection with the convention will include the President's Luncheon, Banquet, and Cocktail Party.

RMA/IRE Meeting Hears of Simplified Television Study

Recent engineering studies which point the way to simplified television sets and improvements in television picture contrast were revealed at the November meeting of members of the engineering department of the Radio Manufacturers Association and The Institute of Radio Engineers at the Hotel Syracuse by W. B. Whalley (A'39) and A. E. Martin (S'40–A'43–M'47) of the Physics Laboratories of Sylvania Electric Products Inc., Bayside, L. I., N. Y.

The study of television receiver simplification, according to Mr. Whalley, "commenced with an analysis of basic television receiver requirements capable of yielding high-quality performance. It revealed that simpler and therefore less expensive circuits will give high-quality television reception. With fewer components there should be fewer failures and less need for servicing. One way in which simplification is possible is through the development of multipurpose tubes in which three distinct functions are provided simultaneously in one tube in contrast to three tubes used in current designs."

Mr. Martin delivered a paper entitled "An Evaluation of Television Viewing Tubes," which was co-authored by R. M. Bowie (A'34–M'37–SM'43–F'48), manager of the Sylvania Physics Laboratories, in which was stated "tint or color in filters is of questionable value. The illumination industry would have removed undesirable portions of the visible spectrum long ago, if they existed. The only real advantage of viewing filters is their ability to alleviate the loss of contrast caused by ambient light in the room, halation, reflection from the back of the safety window and 'hot spots' due to reflections from the curved face of the viewing tube.

"Improvement in contrast when filters are used," Martin continued, "is due to the fact that the extraneous light must pass through filter material more times, and for a greater distance, than the direct light from the picture.

"In tests performed in the Sylvania Physics Laboratories we used twenty people not associated with television research and two TV sets operated with the same fixed value of screen brightness. These people were asked to judge between normal picture tubes and special tubes in which the light transmission was reduced to between 43 per cent and 73 per cent of normal. The surrounding illumination was about the average required for reading."

These tests, according to Messrs. Martin and Bowie, indicated that "dark-faced" tubes were preferable to clear tubes, the preferred light transmission being between 50 per cent and 60 per cent, and that this range is for a level of picture tube brightness that many television sets currently marketed cannot provide.

Mr. Martin stated that a Committee of the Joint Electron Tube Council has adopted an industry recommendation for glass tubes which recognizes the apparent desirability of reducing the light transmission of television picture-tube faces.

IRE Board of Directors Vote Regional Changes

Regional Boundary changes were approved by the Board of Directors of The Institute of Radio Engineers at a regular meeting on Wednesday, September 14, 1950, at headquarters, 1 East 79 Street, New York City. The establishment of two additional sections, Akron and Schenectady, also was voted by the session.

It was agreed that the counties of Franklin, Herkimer, Jefferson, Lewis, and St. Lawrence in upper New York State should be transferred from the New York Section (Region 2) to the Syracuse Section (Region 4).

Other boundary changes were voted as follows: transference from Region 4 to Region 5, and the addition to the Milwaukee Section (transferred from the Detroit Section), of the Upper Peninsula of the State of Michigan, consisting of the counties of Gogebic, Ontonagon, Iron, Houghton, Keweenaw, Braga, Dickinson, Marquette, Menominee, Delta, Alger, Schoolcraft, Mackinac, Luce, and Chippewa; transference of the New Mexico Section from Region 6 to Region 7.

The Board of Directors decided that if and when Sections are formed in Alaska and the Hawaiian Islands, these Sections shall be added to Region 7.

Institute of Radio Engineers Announces Officers, Directors

Raymond F. Guy (A'25–M'31–F'39) and Sir Robert Watson-Watt (SM'45–F'47) have been elected President and Vice-President, respectively, of The Institute of Radio Engineers for 1950, according to an announcement made on Wednesday, November 16, by the Board of Directors of the Institute.

Serving with them for the 1950–1951 term are William R. Hewlett (S'35–A'38–SM'47–F'48) and James W. McRae (A'37–F'47) as Directors-at-Large. Elected as Regional Directors are Herbert J. Reich (A'26–M'41–SM'43–F'49), Region 1; Ferdinand Hamburger, Jr. (A'32–M'39–SM'43), Region 3; John D. Reid (A'41–M'44–SM'46–F'49), Region 5; Austin V. Eastman (A'23–M'32–F'41), Region 7.

Technical Secretary Reports Professional Groups' Status

There were eight active IRE Professional Groups in existence as of November 1, 1949, according to a report on the status of the groups by the Technical Secretary of the IRE to the Committee on Professional Groups. As of that date the following groups had active status: Antennas and Propagation; Audio; Broadcast and Television Receivers; Broadcast Transmission Systems; Circuit Theory; Nuclear Science; Quality Control; and Vehicular and Railroad Radio Communications.

The report further stated nine potential Professional Groups as follows: Electronic Miniaturization; Telemetering; Standards Engineering; Airborne Electronics; Electronic Computers; Electronic Instrumentation; Measurements and Instrumentation; Basic Sciences; and Geophysical Research.

A successful joint IRE/AIEE Conference on Electronic Instrumentation in Nucleonics and Medicine was held on October 31, November 1 and 2, with the Professional Group on Nuclear Science taking an active part. Total registration was 733 and approximately 370 orders for Proceedings of the Symposium have been received to date.

There were approximately 340 people at each technical session when the Antennas and Propagation Group held a meeting in co-operation with URSI in Washington on October 31, November 1 and 2. The final registration figure is expected to be considerably above this. This Professional Group has an enrollment of 509.

The following Professional Groups sponsored sessions at the RMA Radio Fall Meeting in Syracuse October 31, November 1 and 2: Audio, Broadcast and Television Receivers, and Quality Control.

The Audio Group has 100 enrolled members. Increased activity is expected from this Group's new schedule of operation. The Quality Control Group has just started its membership drive and intensive activity is expected from this aggressive organization.

The following Professional Groups will actively participate in the IRE 1950 National Convention by sponsoring technical sessions in co-operation with the Technical Program Committee: Audio, Broadcast and Television Receivers, Circuit Theory, Nuclear Science, and Quality Control.

It was reported that several instances have occurred where such interest in Professional meetings has been stimulated that Sections have requested permission of Professional Groups to send official Section delegates to Professional Group meetings.

Technical Committee Notes

Under the Chairmanship of Professor J. G. Brainerd, the **Standards Committee** met on October 13. The New Measurements Technical Committee has been renamed the **Committee on Measurements and Instrumentation** with Professor Ernst Weber as Chairman. It will supplement the work of ASA Sectional Committee C-42, and will feed the new ASA Committee on Electronic Instruments to be sponsored by IRE. **The Committee on Measurements and Instrumentation** will sponsor a Symposium on Components at the 1950 National Convention. The scope of the Industrial Electronics Committee was approved . . . . The **Video Techniques Committee** held a meeting on October 7. J. E. Keister, Chairman, will write to the Chairman of the Society of Television Engineers in San Francisco, and ask that one or more members be nominated to service on the Video Techniques Subcommittees. Co-ordination with the Television Broadcasters Association will be affected in the same manner. Dr. Garman presented the first paper of his group's series of tutorial papers, "Comparative Factors of Picture Resolution in Television and Film Industries," by H. J. Schlafy. Upon approval of the main committee, the paper will be presented to the IRE Papers Review Committee for publication consideration in the PROCEEDINGS. The second paper will be written by George Gordon and will deal with characteristics of film for recording. The third paper, "The Characteristics of Cathode-Ray Tubes in Relation to Video Recording," has been assigned to H. Milholland's group and will be written by Messrs. Bentley and Hoghland. The fourth paper, "Some Dimensional Aspects of Apparatus Used in Video Recording," has also been assigned. **The Subcommittee on Video Systems and Components**; Methods of Measurement, has **completed** the following **standards**: (1) Measurement of Resolution; (2) Slope Duration and Timing of Synchronizing Pulses; (3) Measurements of Timing on Video Switching Systems. These Standards will be submitted for approval at the November meeting of the Standards Committee . . . . **The Circuits Committee** meeting October 20, under the Chairmanship of W. N. Tuttle, made plans for a Symposium on Circuit Theory to be held at the 1950 National Convention. A Nominating Committee for the Administrative Committee of the **Professional Group on Circuit Theory** was appointed. The members are Professor Brainerd, Chairman, Dr. Dietzold, and Professor Weber. . . . A meeting of the **Wave Propagation Committee** was held on October 19, under the Chairmanship of C. R. Burrows. H. O. Peterson, Chairman of the Subcommittee on Standards and Practices, reported on the activities of his committee. Reports were also made by H. G. Booker, Chairman of the Subcommittee on Theory and Application of Tropospheric Propagation; Chairman H. W. Wells, of the Subcommittee on Theory and Application of Tropospheric Propagation; and Chairman H. W. Wells of the Subcommittee on Theory and Application of Ionospheric Propagation. A report of the Subcommittee on Publications, Definitions, Proposed Standards and Annual Review was given by Chairman A. G. Fox, who has found it necessary to be relieved of his duties. He will be succeeded by George Sinclair. . . . The **Industrial Electronics Committee**, meeting on October 14, with D. E. Watts, Chairman, heard reports of activities of three Subcommittees. . . . A meeting of the **Planning Group for the Joint IRE/AIEE Symposium** on Minimization in Electronic Equipment was held at IRE headquarters on October 10 when it was recommended that the foregoing title be adopted. It was also recommended that the Symposium be sponsored jointly by IRE and AIEE. It will be held in collaboration with RMA and the Bureau of Standards. The Armed Forces will be well represented. In order to make the Symposium readily accessible to Government and Armed Forces Personnel, it is recommended that it be held in Washington, D. C., the second week in May, 1950. . . . **The RMA-IRE Co-ordination Committee** met October 25 and a memo from TBA entitled "Recommendations of the Television Broadcasters Association Concerning Standardization in Television" was discussed by the Committee and the responsibility for undertaking standardization within their respective spheres was assigned to RMA and IRE. To further co-ordination of RMA and IRE activities, a system was proposed for setting up a liaison between those committees of the two organizations having common interests. . . . The **Receivers Committee**, R. F. Shea, Chairman, met October 31, at the Hotel Syracuse during the RMA Radio Fall meeting. . . . The **Federal Communications Commission** held an Engineering **Conference** on Oscillator Radiation on November 1 at Washington, D. C. K. A. Chittick represented the IRE Technical Committee on Receivers and also the Radio Manufacturers Association. C. W. Frick of General Electric Company represented the **IRE Industrial Electronics Committee**, and **Measurements and Instrumentation Committee.** E. K. Jett and James Veatch attended this Conference as Special Consultants to the Joint Technical Advisory Committee. This Conference was particularly well attended by representatives from the electrical industry and power groups. . . . The **Joint Meeting of** the **International Scientific Radio Union and IRE** was held on October 21, November 1 and 2, at the National Academy of Sciences, Washington, D. C. . . . The Administrative Committee of the **Professional Groups on Antennas and Wave Propagation** held a meeting. The next technical meeting to be sponsored jointly by the IRE Professional Group and URSI will be held on the Pacific Coast in February, 1950, at dates and location to be announced later. . . . The **Standards Co-ordinator** plans to hold meetings of the following Institute Committees and Professional Groups at the 1950 National Convention, March 6 to March 9: Standards Committee; Professional Groups Committee; Professional Groups Chairmen; Measurements and Instrumentation Committee; Television Systems Committee; Research Committee; Facsimile Committee; Piezoelectric Crystals Committee. . . . A meeting of the **Professional Group Chairmen** was called by the IRE Standards Coordinator, Dr. Baker, on October 26, at IRE headquarters. Plans for Group participation in the 1950 National Convention were discussed and the Chairmen of the 1950 National Convention Technical Program Committee were present to give information on times available. Each Group Chairman reported on his Group's activities. . . . The **constitution of** the **Professional Group on Nuclear Science** was **approved** by the Executive Committee at the October 10 meeting. It was ruled that Professional Group constitutions need not go before the Executive Committee, but should be approved by the Executive Secretary and the Standards Coordinator. . . . The **Administrative Committee of** the **Professional Group on Quality-Control** held its first meeting on September 19 to draft By-Laws and a Constitution. Officers were elected and the program for the coming year was proposed.

New York Section of AIEE Slates Network Symposium

"Modern Methods of Network Theory and Their Application" will be the subject of a Symposium to be held under auspices of the Basic Science Division of the New York Section of the American Institute of Electrical Engineers. Lectures will be given on January 11, February 15, March 15, and April 5 at 7:00 p.m. at the Consolidated Edison Auditorium, 2 Irving Place, New York, N. Y. The series began in November.

Authorities who will speak at the symposium have been announced as follows: Ph. Le Corbeiller, R. L. Dietzold, S. Darlington, R. M. Foster, and E. A. Guillemin.

The Engineers Joint Council Stresses Professional Unity

The Engineers Joint Council held a meeting October 20 to discuss means by which the engineering organizations of the country could co-operate in increasing the unity of the engineering profession. Representatives of 14 societies were present. The date of the next meeting was set tentatively for January 27.

There was general agreement among the members of the group that some steps looking toward organizing for increased unity of the engineering profession would be desirable, and that such an organization should speak for the profession as a whole. It was felt that the first approach to this subject would best be made through considering methods for closer co-operation between the existing organizations and for modifying, and perhaps grouping, some of these organizations, rather than to consider the establishment of an entirely new, additional organization.

Appointment was made of a Planning Committee, charged with the responsibility for preparing statements regarding various alternative plans for achieving this aim and reporting back to the main group.

The committee includes: B. E. Shackelford, Chairman, The Institute of Radio Engineers; L. W. Bass, American Institute of Chemical Engineers; Edgar J. Kates, The American Society of Mechanical Engineers; T. G. LeClair, American Institute of Electrical Engineers; Dean Thorndike Saville, American Society of Engineering Education; Alex Van Praag, Jr., National Society of Professional Engineers; R. E. Dougherty, ex-officio, Engineers Joint Council; and H. S. Osborne, Secretary.

Swiss Committee Discusses International TV Standards

At their October 18, 1949, meeting, the Swiss Television Committee resumed discussion on television standards with respect to an international plan. It was felt that an image of 625 lines would tend to become more and more a worldwide standard. It was also agreed that in order to have as complete a compatability with existing American standards as possible, a bandwidth of 4.25 Mc seemed the most favorable.

After the discussion, the industrial research section of the Institute of Technical Physics, EPF, demonstrated the definition obtainable with various systems, including systems using 405 and 819 lines.

IRE Cincinnati Section Slates Television Spring Conference

"Television" will be the theme of the Fourth Annual Spring Technical Conference to be sponsored by the IRE Cincinnati Section, on Saturday, April 29, at the Engineering Society headquarters in Cincinnati.

Sessions will be held morning and afternoon, and exhibits will be displayed. The conference will be concluded with a banquet that evening.

Path Rainfall Affects The Attenuation of Microwaves

Microwave radio signals decrease in intensity as they travel through the earth's atmosphere because of absorption and scattering by oxygen, water vapor, or precipitation, according to a report from the National Bureau of Standards.

The attenuation, the report states, increases sharply as microwave frequencies above 10,000 megacycles, and quantitative information on this effect is more important in the selection and allocation of microwave radio frequencies. Annual probability curves for the expected duration and magnitude of atmospheric attenuation at microwave frequencies for both 1-kilometer and 50-kilometer path lengths have now been obtained by Howard E. Bussey at the National Bureau of Standards. These attenuation statistics have been derived from meteorological records, using accepted theoretical and experimental coefficients for converting rainfall values into radio attenuation values.

NBS Publishes Booklet Of HF Voltage Measurements

An up-to-date presentation of fundamental principles and techniques used in high-frequency voltage measurements is given in a new booklet, "High-Frequency Voltage Measurements," published by the National Bureau of Standards and now available from the U. S. Government Printing Office.

The booklet deals primarily with measurements at frequencies in the upper audio- and radio-frequency ranges, including part of the ultra-high frequency range. The following measurements are discussed: high-precision methods based on dc measurements, moderate-precision methods, including thermionic and other rectifiers, pulse-peak voltage measurements, and miscellaneous methods.

Calendar of COMING EVENTS

AIEE Winter General Meeting, New York, N. Y. January 30–February 3, 1950

1950 IRE National Convention, New York, N. Y., March 6–9

Fourth Annual Spring Technical Conference, Cincinnati Section, IRE, April 29, Cincinnati, Ohio

1950 IRE Technical Conference, Dayton, Ohio, May 3–5

Armed Forces Communications Association 1950 Annual Meeting, May 12, New York City, and Long Island City: May 13, Signal Corps Center, Fort Monmouth, N. J.

Radio Fall Meeting, Syracuse, N. Y., October 30–November 1

New York FM Homes Outnumber AM Homes in 26 Other States

There are more FM equipped homes in the metropolitan New York area than there are AM equipped homes in 26 states, according to a survey completed by the FM Association. The study showed that there are more than 520,000 homes in the New York metropolitan area equipped to receive FM programs.

Among the 26 states listed by the FMA where New York FM homes outnumber AM homes, the lowest was Vermont with approximately 90,000 dwellings equipped with AM sets, and the highest was Mississippi with 515,369 AM homes.

New Electronic Accelerator Speeds Automatic Elevators

Sensitive vacuum-tube devices to accelerate the automatic dispatching of high-speed elevators have been developed by engineers of the Westinghouse Elevator Division, Jersey City, N. J. These electronic circuits transmit thousands of split-second impulses every hour, directing and spacing elevator cars so that they will be where they are needed at the right time.

These lightning-fast vacuum-tube devices have been incorporated into the Westinghouse Selectomatic automatic elevator control system, making it an electronic dispatcher for elevators. The Selectomatic system automatically integrates cars, floors, and push-button calls into a smooth flow of service to all floors, even during rush hours.

First installation of the electronic control system will be made in the Merchants Exchange, Memphis, Tenn., and the projected 39-story Mellon-U. S. Steel skyscraper at Pittsburgh, Pa.

ARRL Releases Film on Television Interference

Release of a motion picture on the subject of television interference has been announced by the American Radio Relay League, national organization of radio amateurs, with headquarters at West Hartford, Conn.

The film, photographed, edited, and released entirely through amateur efforts, depicts technical corrective measures which may be taken by the amateur whose neighbors experience trouble from the operation of his private shortwave transmitter. In addition, the film clearly illustrates picture interference from other sources by typical interference patterns, and suggests steps that may be taken to clear up the trouble in many such cases.

Distribution of the film is to be through the more than 600 local amateur radio clubs throughout the country which are affiliated with the parent organization, with early bookings scheduled in congested areas where television interference already constitutes a problem.

NYU Plans Conference on Nuclear Technology

New York University in co-operation with the Atomic Energy Commission will hold for the first time a three-day conference on Industrial and Safety Problems of Nuclear Technology.

Scheduled to be conducted on January 10, 11, and 12, at New York University, the conference will be the first of its kind in the area of industrial applications of nuclear technology.

Planned by Assistant Professor Sidney G. Roth, director of technical courses at the Division and Dr. Walter Cutter of the Safety Center, the conference will attempt to meet the needs of industrial firms, as well as technicians who want to know more about how atomic energy can be applied to peacetime uses.

The first day's discussion will be concerned with problems created by the production of radio-active materials with special emphasis on the hazards associated with radiation.

The use of isotopes and the requirements of a radio-chemical laboratory for safe handling of isotopes will be considered on the second day.

The third and final day of the conference will be devoted to questions of public health-water pollution, sewage disposal, etc., which might arise from the establishment of radio-chemical laboratories and Atomic Energy Commission installations. Important questions of insurance will also be discussed by leading investigators in the field.

General Electric Develops Lightweight X-Ray Machine

Designed especially for use under strenuous military conditions anywhere in the world, two radically new, lightweight, "knock-down" X-ray machines have been demonstrated at the National Naval Medical Center at Bethesda, Md.

Engineers of the General Electric X-Ray Corporation at Milwaukee, Wis., developed the machines which were designed to specifications supplied jointly by the Medical Research and Development board and the National Bureau of Standards.

They will also undergo severe field testing under simulated combat conditions by U. S. and Allied Military forces. Although of identical design, the two machines differ in that one has a structure made primarily of sheet steel, while the other is made primarily of sheet aluminum.

The steel unit weighs approximately 1,600 pounds, and the aluminum unit about 1,100 pounds, compared with approximately 3,000 pounds for a similar X-ray machine available for conventional use.

In addition to their light weight, combined with durability under stress, the new units may be easily serviced, and may be assembled or disassembled by relatively unskilled personnel in a period of time far shorter than has ever before been possible on complete X-ray examination machines.

Industrial Engineering Notes[1]

Television News

FCC has amended its **rules governing diathermy** and other nonbroadcast apparatus to relieve these operators from being required to eliminate interference to television and other receivers arising from direct intermediate pickup of emissions from this type of equipment. The new rules, which have the effect of putting the initial responsibility of correcting TV receivers on set manufacturers, went into effect December 1, 1949. . . . **The FCC's proposed report** on the interconnection features of the AT&T television relay lines **was opposed in petitions** filed **by the Allen B. DuMont** Laboratories **and Western Union.** The AT&T also filed an exception to the proposed report stressing the importance and size of its investment in television relay facilities. DuMont objected to the FCC report because it felt the Bell System program for many years in the future would be inadequate to provide any inter-city TV network connection to many cities to which vhf channels have been allocated, and to more communities for which uhf channels are proposed. . . . **September television receiver production broke** all previous **industry records,** according to RMA tabulations, as trade reports indicated sales of TV sets are continuing to tax the industry's capacity to produce. TV set output by RMA member-companies in September reached 224,532, far ahead of any previous month, with estimated industry production about 265,000. September's record production brought the total television receivers reported in 1949 by RMA members to 1,402,-840 . . . more than three times the number reported for the corresponding period in 1948. Total industry output of television receivers since the war has passed 2,750,000, according to unofficial estimates. FM and FM-AM radio set production in September, not counting TV sets equipped with FM, also rose to 70,936, as against 64,179 in August. However, 43,436 TV receivers included FM reception facilities. . . . Indices of **radio part sales,** as reported by members of the RMA Parts Division, **were higher in September** than in the corresponding month of 1948, according to tabulations by RMA. . . . **Internal Revenue Bureau headquarters has advised RMA** that **radio loudspeakers,** whether with or without coupling devices (output transformers), **are taxable** under the radio excise tax law. This reverses a previous decision that speakers not having a coupling device, or speakers having a coupling device but not "suitable for use" with a radio receiver, were tax exempt. . . . **The FCC has amended its Standards of Good Engineering Practices Concerning FM Broadcast Stations** to establish the ratio of desired to undesired signal intensities for stations separated by 400 and 600 kc. It amended its rules to remove certain requirements of minimum coverage of class-A and class-B stations. . . . The FCC has advised a Virginia Congressman that private business, not the Government, should provide television service in small communities. The Commission turned down a suggestion, which was relayed through Rep. Burr P. Harrison (D., Va.) that the Government establish television stations to cover areas not served by commercial transmitters. However, the FCC said it was "fully aware of the need for sufficient television channels to provide a nation-wide competitive system that would serve the entire country and not just those fortunate enough to reside in large metropolitan areas." . . . **The FCC** this week **received a proposal** that 250,000 amateurs or "fan experimenters" be authorized to assist in "trial and error" testing of proposed color television systems. The Arco Electronics, distributors of electronic components, requested the Commission "to provide color telecasts in major metropolitan markets for a minimum of hours each day so that interested individuals could test their signals." Arco said that it and several hundred other qualified firms are prepared to provide essential parts and components, either in separate form or in knock-down kits for the use of "fan experimenters . . . ." **At the end of October there were 90 television stations** on the air **and 22 construction permits** outstanding, according to FCC records . . . .

The FCC has announced a revised schedule of the color television hearings extending the current inquiry well into 1950 and making it unlikely that any decision can be reached until late spring or early summer. While denying technically an RCA petition, which was supported by RMA, for a postponement of a comparative color versus black-and-white demonstration scheduled for November 14, the FCC in fact extended the proceedings into next year and agreed to hold additional comparative demonstrations between proponents of the three color systems on February 8 . . . . **In revising the hearing schedule, the FCC also postponed cross-examination from December 5, until the middle of February.** The FCC also cancelled a demonstration of the CTI system which was to have been held in San Francisco the week of November 28. CTI will demonstrate its equipment in Washington on February 6 and will participate in the second comparative demonstration on February 8, the FCC said . . . . **Committee No. 4 of the RMA Color Television Committee** this week **submitted a final report to the full committee** headed by W. R. G. Baker, Director of the RMA Engineering Department. Committee No. 4 was charged with detailing the specific nature of field tests which should be conducted to determine the effects of any change of the present television transmission standards. The group has held three further meetings since the FCC requested additional information during the RMA presentation early in the current color television hearings. T. T. Goldsmith, Jr., is Chairman of the Committee . . . . **RMA has supplied the FCC with additional information in support of the Association's position that any color television system should be thoroughly field-tested and proven before standards are adopted.** "Specifically, a field test serves a number of useful purposes," the report

[1] The data on which these Notes are based were selected, by permission, from *Industry Reports*, issues of October 21, October 28, November 4, November 10, published by the Radio Manufacturer's Association, whose helpful attitude is gladly acknowledged.

said. "First, it demonstrates whether or not the new system will give satisfactory performance under the variety of conditions found in the field. Even if an effort were made to simulate field conditions in laboratory tests, the field tests would be necessary as a check that the simulation was adequate and complete." The report pointed out that field trials of a new system "may disclose a natural phenomenon which had but small effect on older systems but which was important to the performance of the new system." It is not likely that every field condition will be reproduced in a laboratory trial, and consequently field tests must be made, the Committee said. "Field tests provide the opportunity to obtain a 'customer' or nontechnical appraisal of a system or new product under home conditions," the report pointed out, adding: "Such an appraisal is necessary to determine whether or not engineers have over-emphasized certain advantages or limitations and underestimated others . . . ." Direct testimony in the FCC color television hearing, at the end of seven weeks, had filled 5,738 pages of the record with approximately 1.1 million words through November 8, not including exhibits. It is now recessed until the resumption in February, 1950.

Radio and Television News Abroad

Plans of a tentative or preliminary **nature were formulated by an industry-government group** during a November meeting called by the State Department **for discussion of International television problems.** The consensus was that this government should send a technical expert to assist the Government of Uruguay in establishing a television system in that country. RMA learned that the FCC said it would make one of its television experts available to assist Uruguay. It was also suggested that JTAC Chairman Donald G. Fink, who attended the Zurich conference as industry representative and adviser to the U.S. Delegation, again be sponsored by RMA. The group recommended that the demonstrations of American television before members of the CCIR Study Group No. 11 be postponed due to a conflict with the FCC color television hearings in the spring. It was suggested that the Study Group visit this country following demonstrations in the United Kingdom and the Netherlands. A general discussion was also held on the best method of furthering the adoption of U.S. television standards in Latin America . . . . The **second television broadcasting station in England** was scheduled to begin operations on December 17 at Birmingham. At the end of July there were 11,783,854 radio licenses and 148,618 television licenses outstanding in Britain . . . . The **recent television exhibition at Milan, Italy,** on September 10–19, **attracted 100,000 persons and 40 exhibitors** with products valued at $1,739,-000, according to information received by the U. S. Department of Commerce. Eighteen U.S. television equipment manufacturers exhibited their products at the "First International Television Exhibition and Technical Convention." Seventeen other manufacturers represented Italy, while France and Britain were represented by three concerns each . . . . **Production of radio apparatus in Finland** during 1948 totaled 900,000,000 marks. Receivers accounted for 70 per cent of the total and marine apparatus accounted for 20 per cent, while other radio equipment amounted to 10 per cent . . . . **There were an estimated 54,000 radio receivers in use in Bolivia** the first of 1949 compared with 50,000 in the corresponding period in 1948. The U.S. is the principal source of supply . . . . **Manufacturers in Belgium expect to produce an estimated 120,000 radio receivers** this year. Approximately 150,000 sets were produced in 1947 and 1948 . . . . **The Canadian government plans to appropriate $4,500,000 to** "help the CBC meet the initial capital costs of introducing television in Canada and to erect the first two TV stations" at Montreal and Toronto. The money is due to be repaid by funds obtained from the doubling of the radio license fee from $2.50 to $5.00 per year. This fee would not cover television sets for which owners would be required to obtain a supplementary license of possibly $10 per year after television has been established . . . . **Sales of radio receivers in Canada increased in July,** compared with the corresponding month of 1948. July sales totaled 42,756 units valued at $2,846,958, compared with 20,334 units valued at $1,-290,636 in July of last year. Sales during the first seven months of this year totaled 346,991 units valued at $24,710,190 as against 225,733 units valued at $20,109,467 in the same 1948 period. Canadian imports of radio sets in July totaled 6,454 units valued at $208,498, while exports amounted to 22,332 sets valued at $876,544. Imports of radio receiving tubes totaled 95,893 units valued at $78,969 and tube parts imported aggregated $24,197 during the month of July. Canadian production of radio tubes amounted to $338,527 units valued at $192,210.

Radiation Detection Device In Production for U. S. Army Use

Production has started on a new type of radiation detecting and measuring device suitable for use by both military and civil defense organizations, according to the Department of the Army announcement which said the invention is called a "radiac set" and used standardized parts.

Intentionally made less sensitive than the Geiger counter, the new instrument is designed to detect and measure relatively large concentrations of radiation, such as would result from an atomic bomb blast, the Army said. Specifications were drawn by the Signal Corps.

The announcement states the Atomic Energy Commission, the Navy, and Air Force are developing a variety of instruments for radiation detection to meet their particular needs. The over-all program among the military is co-ordinated by the Armed Forces Special Weapons Project.

The new radiac set is being produced by the Instrument Division of the Kelley-Koett Manufacturing Co. of Covington, Ky., under a Signal Corps contract signed last spring. The Signal Corps also has under production, by El-Tronics, Inc., of Philadelphia, a rugged Geiger counter likewise using all standardized parts.

Bureau of Standards Develops Plates of Ceramic Dielectrics

Because of rapid developments in the electronics field which have introduced operating conditions too severe for electrical insulating materials commonly used as dielectrics, the National Bureau of Standards is investigating the properties of ceramic dielectrics. The Bureau, according to a report, has fabricated "ceramics in the form of thin plates comparable in thickness to that of paper and mica." Ceramic dielectrics are applicable to electronic signal devices not only for the armed services, but also for the commercial production of radio, radar, and television sets and hearing aids, the Bureau said. Further details on the development are contained in Research Paper RP2025, Journal of Research, NBS, September, 1949.

New Radar Set Is Designed To Provide Storm Warnings

A new radar set designed to give weathermen and pilots advance warning of storm areas as far distant as 200 miles from the site of operations has been developed by the Army Signal Corps.

The new equipment is electronically similar to the successful wartime warning radars which told of the approach of enemy aircraft or ships. "However," the Signal Corps said, "it utilizes what was considered a characteristic fault in the earlier radars, namely, their tendency to pick up signals from nearby rainstorms, thereby masking indications from possible enemy targets on the farside of such rainstorms."

Surplus Manual Published by Department of Commerce

A simplified manual for users of the more common type of electronic equipment purchased from Government surplus stocks has just been issued by the Office of Technical Services, U. S. Department of Commerce. The new publication provides prospective purchasers with the basis circuit diagram, parts, values, and voltages of the equipment listed. Copies of PB 98487, "Schematic Manual for Surplus Electronic Equipment," are available from the OTS at $1.00. Orders should be accompanied by check or money order payable to the Treasurer of the United States.

State to Spend 10.5 Million For Anti-Jamming Facilities

The U. S. State Department just before the adjournment of Congress in October was given an appropriation of $10,475,000 to purchase and install anti-jamming facilities in connection with its efforts to penetrate the Soviet Iron Curtain with its Voice of America broadcasts.

IRE People

Alexander Ellett (SM'48) has been elected vice-president in charge of research, according to an announcement from the directors of the Zenith Radio Corporation. Dr. Ellett has headed Zenith's research laboratories since 1946. One of his major contributions has been to Phonevision, a Zenith development which has been in the laboratory since 1931, and which Dr. Ellett made commercially practicable.

During the war Dr. Ellett was head of Division 4 of the NDRC, where he directed the development of the famous V-T proximity fuze for bombs and rockets, the only research project of the war which shared top priority with the atomic bomb.

Dr. Ellett was awarded the President's Medal for Merit in June, 1948, for his development of the proximity fuze and of printed ceramic circuits. Prior to his affiliation with the NDRC, Dr. Ellett was professor of physics at the University of Iowa, where his major research activities were in spectroscopy, atomic and molecular beams, and in nuclear physics. He served two years in the service in World War I.

Britton Chance (M'46–SM'46), Director of the Eldridge Reeves Johnson Foundation for Medical Physics of the University School of Medicine, has been awarded the President's Certificate of Merit "for his important contributions to the development of radar during World War II." President Truman's citation relates how Dr. Chance was responsible for contributions to the development of radar circuits while stationed at the Radiation Laboratory of the Massachusetts Institute of Technology under auspices of OSRD from 1941 to 1945, inclusive. His service at MIT included the following assignments: leader of the precision Components Group, 1942–1945; associate head, receiver components division, 1943; and a member of the Steering Committee, 1943–1944.

During his work with radar Dr. Chance devised improvements of the Roughton and Millikan flow method for the study of rapid reactions to a stage where optical density changes of less than 0.001 could, through high amplifications of a photocell current, be rapidly and accurately recorded.

In 1945–1946 Dr. Chance was a member of the Editorial Board, Radiation Laboratory Series, editing and writing a greater part of the book, "Waveforms," which describes the new electronic circuits used in computation of precise distance for time measurements for radar. Dr. Chance also edited and wrote most of the book, "Electronic Time Measurements."

Dr. Chance was graduated from the University of Pennsylvania in 1935, received the master of science degree the following year, and was granted the Ph.D. degree in physical chemistry in 1940, also from the University of Pennsylvania. Dr. Chance was awarded the Ph.D. degree in biology from Cambridge University in 1943.

He was appointed assistant professor of biophysics at the University of Pennsylvania in 1941, and for a time served as Acting Director of the Johnson Foundation. Immediately following the war he held a postservice Guggenheim Fellowship for two years, and studied at the Medical Nobel Institute in Stockholm and at the Molten Institute in Cambridge.

Albert D. Martin, Jr. (M'45), principal radio engineer, ASF, Army Airways Communication Service, Plant Engineering Agency, Philadelphia, Pa., died recently following a short illness. He assumed the post in 1942 and was responsible for the procurement and distribution of fixed radio equipment and supplies, preparation of equipment specifications and standard engineering practises, and the technical supervision of installation and maintenance activities of field personnel.

From 1934 until 1935, Mr. Martin was the receiving engineer on international radio telegraph and broadcast circuits and RCA Frequency Monitoring Service with RCA Communications, Inc., Riverhead, L. I., N.Y.

Mr. Martin, who was born on March 20, 1909, at Gunther, Texas, received his early education at Bryan High School, Bryan, Texas. He was graduated in 1929 from Texas A. & M. College with the B.S. degree, and earned the electrical engineering degree in 1935 at the same institution.

From 1929 until 1934 Mr. Martin was with the National Bureau of Standards at Washington, D. C., as an electrical engineer, working on the research, development, construction, operation, and maintenance of facilities for transmission of standard radio frequencies. He was elevated to the position of engineer in charge in 1933.

Myron F. Eddy (M'49) has been appointed Director of Training at Cleveland Institute of Radio Electronics, where he will have charge of planning, writing, editing, and revising all of the Institute's home training lesson text material.

Mr. Eddy (Lieutenant, U. S. Navy, Ret.) studied electrical engineering at Kansas State College, and then served during the war as a Naval aviator and communications officer. He was retired in 1929 as the result of an airplane crash, and has been teaching radio and writing on radio subjects since then. He is the author of magazine articles and several books, including "Aircraft Radio," published in 1930, and "Aeronautic Radio," in 1940.

Harvey J. Klumb (A'26) of the Rochester Gas and Electric Company was presented with the annual award plaque of the Radio Fall Meeting sponsored jointly by the Engineering Department of the Radio Manufacturers Association and The Institute of Radio Engineers, according to **Virgil M. Graham,** (A'24–M'27–F'35) chairman of the meeting and director of technical relations for Sylvania Electric Products Inc.

Mr. Graham said that the presentation of the plaque to Mr. Klumb was in recognition "of his twenty years of hard work for the Annual Fall Meeting." The plaque, a handsome gold and cloisonne design, is the ninth to be awarded since the conception of the award by the Fall Meeting Committee "to commemorate contributions outstanding to the radio and electronic art above the call of duty."

The first Fall Meeting plaque was awarded to **W. R. G. Baker** (A'19–F'28) of the General Electric Company in 1941 for his work in the organization and management of the National Television Committee. Last year the award was made to **Dorman D. Israel** (A'23–M'30–F'42) of the Emerson Radio and Phonograph Corporation. Other recipients of the annual award have included: 1942, **L. C. F. Horle** (A'14–M'23–F'25), chief engineer, Radio Manufacturers Association; 1943, **R. A. Hackbusch,** (A'16–M'30–F'37) president, Stromberg Carlson of Canada, Ltd.; 1944, **Keith Henney,** (A'18–M'26–SM'43–F'43), consulting editor, *Electronics* Magazine; 1945, **L. A. DuBridge,** (A'35–F'42), president, California Institute of Technology; 1946, **W. L. Everitt** (A'25–M'29–F'38), head, department of electrical engineering, University of Illinois; 1947, **F. S. Barton** (F'35), director of communications development, British Ministry of Supply.

❖

Walter E. Poor (A'29), Chairman of the Board of Sylvania Electric Products Inc., has been awarded a bronze "Oscar of Industry" trophy from Weston Smith, executive vice-president of Financial World for the best 1948 Annual Report in the electronics-radio industry. More than 4,500 entries were submitted in the ninth of a series of annual report surveys conducted by the publication and originated by Mr. Smith.

Mr. Poor, who was born at Peabody, Mass., was graduated from the Massachusetts Institute of Technology with the B.S. degree in electrical engineering in 1908.

❖

Correction

On page 1300 of the November, 1949, issue of the PROCEEDINGS it was indicated that **Edward W. Davis** (A'40) had been elected an officer of the AIEE. The election to directorship in the AIEE actually was that of **Ernest W. Davis,** chief electrical engineer, Simplex Wire and Cable Company, Cambridge, Mass.

Sections*

Chairman	Section	Secretary
H. R. Hegbar 2145 12th St. Cuyahoga Falls, Ohio	AKRON (4)	H. G. Shively 736 Garfield St. Akron, Ohio
H. I. Metz C.A.A. 84 Marietta St., N.W. Atlanta, Ga.	ATLANTA (6) Jan. 20–Feb. 17	M. S. Alexander 2289 Memorial Dr., S.E. Atlanta, Ga.
E. W. Chapin 2805 Shirley Ave. Baltimore 14, Md.	BALTIMORE (3)	J. V. Lebacqz Johns Hopkins Un. Baltimore, Md.
T. B. Lawrence 1833 Grand Beaumont, Texas	BEAUMONT-PORT ARTHUR (6)	C. B. Trevey 2555 Pierce St. Beaumont, Texas
H. H. Scott Hermon Hosmer Scott, Inc. 385 Putnam Ave. Cambridge 39, Mass.	BOSTON (1)	F. D. Lewis General Radio Co. 275 Massachusetts Ave. Cambridge 39, Mass.
J. P. Arnaud Guemes 827 Vte. Lopez F.C.C.A., Argentina, S.A.	BUENOS AIRES	L. Brandt Uruguay 618 Buenois Aires, Argentina, S.A.
L. P. Haner 75 Koenig Rd. Tonawanda, N. Y.	BUFFALO-NIAGARA (4) Jan. 18–Feb. 17	K. R. Wendt Colonial Radio Corp. 1280 Main St. Buffalo 9, N. Y.
M. S. Smith 1701 10th St. Marion, Iowa	CEDAR RAPIDS (5)	V. R. Hudek Collins Radio Co. Cedar Rapids, Iowa
E. H. Schulz Elec. Engr. Dept. Armour Research Found. Chicago, Ill.	CHICAGO (5) Jan. 20–Feb. 17	L. H. Clardy Research Labs. Swift & Co., U. S. Yards Chicago 9, Ill.
F. W. King 6249 Banning Rd. Cincinnati 24, Ohio	CINCINNATI (5) Jan. 17–Feb. 14	J. P. Quitter 509 Missouri Ave. Cincinnati 20, Ohio
J. F. Dobosy 31748 Lake Rd. Avon Lake, Ohio	CLEVELAND (4) Jan. 26–Feb. 23	T. B. Friedman 2909 Washington Blvd. Cleveland Heights 18, Ohio
R. B. Jacques 226 W. Como Ave. Columbus, Ohio	COLUMBUS (4) Jan. 20–Feb. 17	S. N. Friedman 144 N. Edgevale Columbus, Ohio
Lawrence Grew S. N. E. Telephone Co. New Haven, Conn.	CONNECTICUT VALLEY (1) Jan. 19–Feb. 26	J. E. Merrill 16 Granada Terr. New London, Conn.
A. S. Levelle 801 Telephone Bldg. Dallas 2, Texas	DALLAS-FORT WORTH (6)	E. A. Hegar 802 Telephone Bldg. Dallas 2, Texas
H. E. Ruble 3011 Athens Ave. Dayton 6, Ohio	DAYTON (5)	G. H. Arenstein 1224 Windsor Drive Dayton 7, Ohio
T. G. Morrissey Radio Station KFEL Albany Hotel Denver, Colo.	DENVER (5)	Hubert Sharp Box 960 Denver 1, Colo.
F. E. Bartlett Radio Station KSO Old Colony Bldg. Des Moines 9, Iowa	DES MOINES-AMES (5)	O. A. Tennant 1408 Walnut St. Des Moines, Iowa
C. F. Kocher 17186 Sioux Rd. Detroit 24, Mich.	DETROIT (4) Jan. 20–Feb. 17	P. L. Gundy 519 N. Wilson Royal Oak, Mich.
R. W. Slinkman Sylvania Elec. Prods. Emporium, Pa.	EMPORIUM (4)	T. M. Woodward 203 E. Fifth St. Emporium, Pa.
H. W. G. Salinger 2527 Hoagland Ave. Ft. Wayne 6, Ind.	FORT WAYNE (5)	J. F. Conway 4610 Plaza Dr. Ft. Wayne, Ind.
C. R. Wischmeyer 808 N. Rice Ave. Bellaire, Texas	HOUSTON (6)	Wayne Phelps 26 N. Wynden St. Houston 6, Texas
E. H. Pulliam 931 N. Parker Ave. Indianapolis 1, Ind.	INDIANAPOLIS (5)	J. H. Schult Indianapolis Elec. School 312 E. Washington St. Indianapolis 4, Ind.
E. R. Toporeck Naval Ordnance Test Sta. Inyokern, Calif.	INYOKERN (7)	R. W. Johnson 303 B. Langley China Lake, Calif.
C. F. Heister Fed. Com. Comm. 838 U. S. Court House Kansas City 6, Mo.	KANSAS CITY (5)	Mrs. G. L. Curtis 6005 El Monte Mission, Kan.
G. L. Foster Sparton of Canada London, Ont., Canada	LONDON, ONTARIO (8)	G. R. Hosker Richards-Wilcox London, Ont., Can
R. L. Sink Consolidated Eng. Co. 620 N. Lake Ave. Pasadena 4, Calif.	LOS ANGELES (7)	W. G. Hodson Northrop Aircraft Telecommunications Sec. Hawthorne, Calif.
D. C. Summerford Radio Station WKLO Henry Clay Hotel Louisville, Ky.	LOUISVILLE (5)	R. B. McGregor 2100 Confederate Pl. Louisville, Ky.
E. J. Limpel A. O. Smith Corp. 3533 N. 27 St. Milwaukee 1, Wis.	MILWAUKEE (5)	W. H. Elliot 3564 N. Murray Ave. Milwaukee 11, Wis.
A. B. Oxley R.C.A. Victor Co. 1001 Lenoir St. Montreal, P.Q. Canada	MONTREAL, QUEBEC (8) Jan. 11–Feb. 8	H. A. Audet Canadian Broadcasting Corp. 1231 St. Catherine St. Montreal, Que, Can.
C. W. Carnahan 3169—41 Place Sandia Base Branch Albuquerque, N. M.	NEW MEXICO (7)	T. S. Church 3079 Q 34th St. Sandia Base Branch Albuquerque, N. M.
H. F. Dart 33 Burnett St. Glen Ridge, N. J.	NEW YORK (2)	Earl Schoenfield W. L. Maxson Corp. 460 W. 34th St. New York 1, N. Y.
J. T. Orth 4101 Fort Ave. Lynchburg, Va.	NORTH CAROLINA-VIRGINIA (3)	C. E. Hastings 117 Hampton Roads Ave. Hampton, Va.
M. W. Bullock Capital Broadcasting Co. 501 Federal Securities Bldg. Lincoln 8, Neb.	OMAHA-LINCOLN (5)	B. L. Dunbar Radio Station WOW Omaha, Neb.
A. W. Y. Des Brisay 240 Clemow Ave. Ottawa, Ont., Canada	OTTAWA, ONTARIO (8) Jan. 19–Feb. 16	A. G. Sheffield 11 Fern Ave. Ottawa, Ont., Canada
J. T. Brothers Philco Radio and Television Tiogo and 'C' Sts. Philadelphia 34, Pa.	PHILADELPHIA (3) Jan. 8–Feb. 3	L. M. Rodgers 400 Wellesley Rd. Philadelphia 19, Pa.
M. Glenn Jarrett 416 Seventh Ave. Pittsburgh 19, Pa.	PITTSBURGH (4) Jan. 9–Feb. 13	W. P. Caywood, Jr. 23 Sandy Creed Rd. Pittsburgh 21, Pa.
A. E. Richmond Box 441 Portland 7, Ore.	PORTLAND (7)	Henry Sturtevant Rt. 6, Box 1160 Portland 1, Ore.
E. W. Herold RCA Laboratories Princeton, N. J.	PRINCETON (3)	W. H. Bliss 300 Western Way Princeton, N. J.
K. J. Gardner 111 East Ave. Rochester 4, N. Y.	ROCHESTER (4) Jan. 19–Feb. 16	Gerrard Mountjoy Stromberg Carlson Co. 100 Carlson Rd. Rochester, N. Y.
N. D. Webster 515 Blackwood N. Sacramento, Calif.	SACRAMENTO (7)	J. R. Miller 3991 3rd Ave. Sacramento, Calif.
L. A. Mollman Union Electric Co. 12 and Locust Sts. St. Louis 1 Mo.	ST. LOUIS (5)	H. G. Wise 1705 N. 48 St. E. St. Louis, Ill.
C. R. Evans Radio Station KSL Salt Lake City, Utah	SALT LAKE (7)	E. C. Madsen Dept. of Elec. Eng. University of Utah Salt Lake City, Utah
C. L. Jeffers Radio Station WOAI 1031 Navarro St. San Antonio, Texas	SAN ANTONIO (6)	L. K. Jonas 267 E. Mayfield Blvd. San Antonio, Texas
L. G. Trolese U. S. Navy Electronics Lab. San Diego 52, Calif.	SAN DIEGO (7)	S. H. Sessions U. S. Navy Electronics Lab. San Diego 52, Calif.
W. R. Hewlett 395 Page Mill Rd. Palo Alto, Calif.	SAN FRANCISCO (7)	J. R. Whinnery Elec. Engr. Dept. University of Calif. Berkeley, Calif.

* Numerals in parentheses following Section designate Region number.

Sections

Chairman		Secretary
J. M. Paterson 2009 Nipsic Bremerton, Wash.	SEATTLE (7) Jan. 12–Feb. 9	J. E. Hogg General Electric Co. 710 Second Ave. Seattle 1, Wash.
R. H. Williamson 161 Parkway Dr. Syracuse, N. Y.	SYRACUSE (4)	S. E. Clements Dept. of Elec. Engr. Syracuse University Syracuse, N. Y.
A. M. Okun 344 Boston Pl. Toledo 10, Ohio	TOLEDO (4)	R. G. Larson 2647 Scottwood Ave. Toledo 10, Ohio
C. Graydon Lloyd Canadian General Electric Co., Ltd. 212 King St., W. Toronto, Ont., Canada	TORONTO, ONTARIO (8)	Walter Ward Canadian General Electric Co., Ltd. 212 King St., W. Toronto, Ont., Canada
W. G. Pree 2500 W. 66 St. Minneapolis, Minn.	TWIN CITIES (5)	O. A. Schott 4224 Elmer Ave. Minneapolis 16, Minn.
T. J. Carroll National Bureau of Stand. Washington, D. C.	WASHINGTON (3) Jan. 9–Feb. 13	P. DeF. McKeel 9203 Sligo Creek Parkway Silver Spring, Md.
G. C. Larson Westinghouse Elec. Corp. Sunbury, Pa.	WILLIAMSPORT (4) Jan. 4–Feb. 1	R. C. Walker Box 414, Bucknell Univ. Lewisburg, Pa.

SUBSECTIONS

Chairman		Secretary
H. W. Harris 711 Kentucky St. Amarillo, Tex.	AMARILLO-LUBBOCK (6) (Dallas-Ft. Worth Subsection)	E. N. Luddy Station KFDA Amarillo, Tex.
F. T. Hall Dept. of Elec. Engr. Pennsylvania St. College State College, Pa.	CENTRE COUNTY (4) (Emporium Subsection)	J. H. Slaton Dept. of Eng. Research Pennsylvania St. College State College, Pa.
A. H. Sievert Canadian Westinghouse Co. Hamilton, Ont., Canada	HAMILTON (8) (Toronto Subsection)	J. H. Pickett Aerovox Canada Ltd. 1551 Barten St. E. Hamilton, Ont., Canada
R. B. Ayer RCA Victor Division New Holland Pike Lancaster, Pa.	LANCASTER (3) (Philadelphia Subsection)	J. L. Quinn RCA Victor Division New Holland Pike Lancaster, Pa.
O. M. Dunning Hazeltine Elec. Corp. 5825 Little Neck Pkwy. Little Neck, L. I., N. Y.	LONG ISLAND (2) (New York Subsection)	David Dettinger Wheeler Labs. 259–09 Northern Blvd. Great Neck, L. I., N. Y.
H. Sherman Watson Labs.-ENRPS- Red Bank, N. J.	MONMOUTH (2) (New York Subsection)	W. L. Rehm Signal Corps Eng. Labs. Rm. 247, Squier Lab. Fort Monmouth, N. J.
N. Young, Jr. F.C.C. Nutley, N. J.	NORTHERN N. J. (2) (New York Subsection)	J. H. Redington Measurements Corp. Boonton, N. J.
A. R. Kahn Electro-Voice, Inc. Buchanan, Mich.	SOUTH BEND (5) (Chicago Subsection) Jan. 19–Feb. 16	A. M. Wiggins Electro-Voice, Inc. Buchanan, Mich.
R. M. Wainwright Elec. Eng. Department University of Illinois Urbana, Ill.	URBANA (5) (Chicago Subsection)	M. H. Crothers Elec. Eng. Department University of Illinois Urbana, Ill.
R. D. Cahoon C.B.C. Winnipeg, Man., Canada	WINNIPEG (8) (Toronto Subsection)	J. R. B. Brown Suite 2 642 St. Marys Rd. Winnipeg, Man., Canada

Books

Handbook of Patents by Harry Aubrey Tolmin, Jr., J. D., Litt.D., LL.D.

Published (1949) by D. Van Nostrand Co., Inc., 250 Fourth Ave., New York, N. Y. 610 pages+153-page appendices+6-page index+vii pages. LXIX figures. 6×9. $9.00.

The author's invention is expressed as making it easier for the business man, the inventor, the engineer, and the executive to understand the philosophy upon which the Patent Law is grounded, and the relationship of Patent Law to manufacturing, research, and engineering. The title and the arrangement show that an attempt has been made to provide a handbook, but no handbook is any better than its index. The index appears to be fairly well arranged, but not nearly as complete as might be desired, particularly when compared to the more complete index of the Rules of the Patent Office which covers only 56 pages of subject matter.

Since this book is being reviewed in behalf of the radio engineers and executives, a rather high standard is set, because there is a very definite need for a comprehensive handbook to assist these individuals who encounter matters relating to patents and designs without the benefit of formal education in patent matters. The information that has been presented is easily readable and clearly presented, although in some instances it might be expanded to some extent to serve better the engineer and executive. The importance of design patents to the engineer does not appear to have been given sufficient treatment, and much more could be said about the evidence necessary for the testimony in patent interference matters.

Unfortunately, this book still refers to the old Patent Office Rules which were superseded by new rules March 1, 1949. The old rules and index are presented in full and take up 96 of the 153-page appendix. The remaining pages of the appendix are devoted to the Patent Laws. About 30 pages of the 610 pages of text could have been saved by rearrangement, and other pages could have been saved by eliminating the reproduction of certain cuts which appear in many instances without adequate explanation.

This book appears to be a new volume by an author who is rather a prolific writer on patent subjects. His approach in this instance is quite different from any of his other works, and is commendable in that he appreciates there is a need for a book which can serve a layman. The engineer by referring to the book undoubtedly would have a clearer comprehension of the important part patents play in industry, and hence, can be recommended to those who have a need for a book of this kind.

ALOIS W. GRAF,
135 South La Salle St.
Chicago 3, Ill.

Dynamic Principles of Mechanics by David R. Inglis

Published (1949) by the Blakiston Co., Philadelphia, Pa. 169 pages+4-page index. 64 figures 6×9.

This short book, complete with problems, is suitable for an undergraduate introduction to the study of the dynamics of particles and of rigid bodies.

Vector notation is used consistently, and, where possible, explanations and diagrams are in terms of vectors as directed quantities, rather than in terms of their components in a particular co-ordinate system. Tensors (matrices of inertia) are introduced in connection with rigid bodies. Statics is treated as a special case of dynamics.

The book includes the treatment of coupled oscillators of two and three particles, with an explanation of normal co-ordinates, planetary motion, the equations of motion for rigid bodies, a treatment of gyroscopes and tops, and problems of impact and scattering, both for spheres and for an inverse square law (atomic scattering).

The book is written in good, plain English, and is as easy to read as the contents will allow. The problems are sensible, and are designed to make the student think.

J. R. PIERCE
Bell Telephone Laboratories, Inc.
Murray Hill, N. J.

Abstracts and References

Prepared by the National Physical Laboratory, Teddington, England, Published by Arrangement with the Department of Scientific and Industrial Research, England, and *Wireless Engineer*, London, England

NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications and not to the IRE.

The Institute of Radio Engineers has made arrangements to have these Abstracts and References reprinted on suitable paper, on one side of the sheet only. This makes it possible for subscribers to this special service to cut and mount the individual Abstracts for cataloging or otherwise to file and refer to them. Subscriptions to this special edition will be accepted only from members of the IRE and subscribers to the Proc. I.R.E. at $15.00 per year. The Annual Index to these Abstracts and References, covering those published in the PROC. I.R.E. from February, 1948, through January, 1949, may be obtained for 2s. 8d. postage included from the *Wireless Engineer*, Dorset House, Stamford St., London S. E., England. This index includes a list of the journals abstracted together with the addresses of their publishers.

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number and is not to be confused with the Decimal Classification used by the United States National Bureau of Standards. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger (†) must be regarded as provisional.

ACOUSTICS AND AUDIO FREQUENCIES

53.081.4 3319

A New Frequency Scale for Acoustic Measurements—W. Koenig. (*Bell Lab. Rec.*, vol. 27, pp. 299–301; August, 1949.) A scale which is linear up to 1,000 cps and logarithmic above 1,000 cps, with smooth transition at 1,000 cps.

534.143 3320

On the Principal Possibilities of Electroacoustic Energy Transformation and its Classification—F. A. Fischer. (*Arch. Elek.* (Übertragung), vol. 3, pp. 129–135; July, 1949.) The electrodynamic theory of quasistationary fields gives six fundamentally different possible methods of transforming electrical into mechanical energy in noncrystalline media, three electrical and three magnetic. These are discussed.

534.321.9 3321

Electric-Field Modulation of Ultrasonic Signals in Liquids—A. W. Nolle. (*Jour. Appl. Phys.*, vol. 20, pp. 589–592; June, 1949.) An experiment to determine (a) whether the presence of a periodic transverse electric field produces modulation of the amplitude or phase of a continuous progressive ultrasonic wave train passing through a liquid, and (b) whether the application of an electric field to a polar liquid affects either the compressibility or the viscosity of the liquid through molecular orientation. Phase modulation was observed in some conducting liquids. Amplitude modulation was not found in any liquid. See also 3002 of December (Bonetti).

534.321.9:534.373 3322

Ultrasonic Absorption in Water in the Region of 1 Mc/s.—C. E. Mulders. (*Nature* (London), vol. 164, pp. 347–348; August 27, 1949.) The absorption (of the order of 0.2 db/m) is calculated, by a formula taking friction and radiation into account, from the reverberation time of a smooth aluminum vessel filled with boiled distilled water. See also 932 of May.

534.4 3323

The Analysis and Synthesis of Musical Sounds—A. W. Ladner. (*Electronic Eng.* (London), vol. 21, pp. 379–386; October, 1949.) The synthesis of complex sustained musical tones by means of a series of harmonics is much more difficult than the synthesis of musical transients. This is probably because the waves produced by a musical instrument are not all exactly alike; the second-order amplitude and frequency differences cause a frequency spread which is not revealed by present methods of harmonic analysis. Sounds should therefore be examined also for frequency changes by frequency-discriminating circuits. An acoustical harmonic synthesizer is described.

534.6+534.8 3324

Work of the Laboratory of Technical Acoustics, 1941–1946—F. Ingerslev and K. Nielsen. (*Ingen. Vidensk. Skr.*, no. 2, 128 pp.; 1947. In Danish.) An account of some of the apparatus and methods of measurement developed since the formation of the Laboratory. The layout of lagged and reverberation rooms, is described. Determinations of the radiation pressure, intensity and frequency spectrum of sound in air, measurement of mechanical vibrations in structures, of reverberation times as functions of frequency, and of the distribution of sound in enclosed spaces, determination of sound-absorption coefficients by the tube and sound-chamber methods, insulation against airborne sound and reduction of foot-fall noise in buildings, are discussed. On the electroacoustics side, the measurements described concern principally the frequency and directional characteristics of microphones and loudspeakers. Special attention has been paid by the Laboratory to the measurement of sound-absorption coefficients and details are given of an improved selective amplifier for use with the tube method.

534.61:534.75 3325

On a New Audiometer—G. von Békésy. (*Arch. Elek.* (Übertragung), vol. 1, pp. 13–16; July and August, 1947.) The slider of a potentiometer determining the intensity of the tone provided by an af signal generator is operated by a motor whose rotation in either direction can be controlled by the person whose hearing is to be tested. Intensity is adjusted until a particular tone is inaudible and readjusted until the tone is again audible, the movement of the potentiometer slider being recorded on a drum. This drum is coupled to a variable capacitor by means of which the generator frequency is varied between the limits of 100 cps and 15 kc. The whole frequency range is covered slowly to obtain a zigzag record on the drum. The potentiometer is designed to give intensity steps of 2 db, so that the record is direct-reading for both intensity and frequency. Typical records for hard-of-hearing subjects are given.

534.833.1 3326

Transmission of Reverberant Sound through Single Walls—A. London. (*Bur. Stand. Jour. Res.*, vol. 42, pp. 605–615; June, 1949.) Random-incidence sound-transmission measurements were made on homogeneous walls of plywood and plasterboard. Results were in satisfactory agreement with a modified version of Cremer's theory (1446 of 1943) which postulates that the wall impedance has a resistive component as well as its mass reactance and a stiffness reactance due to flexural waves. The resistance and critical flexure frequency are determined from the experimental data so as to obtain the best fit between theoretical and experimental results; the transmission loss is deduced. The validity of the mass law of sound transmission is discussed. An increase in transmission loss can be obtained by applying a fairly substantial sound-absorbent blanket to the back of a homogeneous wall.

621.395.61/.62:546.431.82 3327

Barium-Titanate Ceramic as an Electromechanical Transducer—Mason. (*See* 3439.)

621.395.625.2 3328

Considerations on Disk Recording and Reproduction—P. H. Werner. (*Tech. Mitt. Schweiz. Telegr.-Teleph. Verw.*, vol. 27, pp. 168–178; August 1, 1949. In French and German.) Detailed discussion of the "Technical Standards and Good Engineering Practices of the National Association of Broadcasters for Electrical Transcriptions and Recording for Radio Broadcasting." This publication was noted in 458 of 1943.

621.395.625.3 3329

Magnetic Recording Technique—D. Roe. (*Wireless World*, vol. 55, pp. 362–364; October, 1949.) Practical notes for the experimenter. See also 3889 of 1944 (Aldous: Ashman) and 2463 of 1946 (Power).

621.395.625.3 3330

Magnetic-Tape Recorder—(*Engineer* (London), vol. 188, pp. 198–199; August 19, 1949.) A high-fidelity speech and music recorder made by the General Electric Co., London. Its 1,000 yd of coated plastic tape give a recording or play-back time of 80 min. The fitting of separate recording, playing, and erasing heads en-

ables continuous monitoring of the recorded program to be carried out.

681.85:531.7 3331
Compliance Meter for Pickups—A. M. Wiggins. (*Electronics*, vol. 22, pp. 94–95; October, 1949.) The stylus point is placed in a V-groove in the edge of a vibrating reed on which is cemented a piezoelectric ceramic strip whose output voltage is measured.

621.395.625.3 3332
Magnetic Recording [Book Review]—S. J. Begun. Publishers: Murray Hill Books, New York, 1949, 242 pp., $5.00. (*Electronics*, vol. 22, pp. 248, 250; October, 1949.) A book that reviews "the subject up to date, separates the chaff from the wheat in the literature, and presents an authoritative discussion of the art. . . . an unusually readable and well organized text."

ANTENNAS AND TRANSMISSION LINES

621.3.09 3333
Calculation of the Impedance and Attenuation of High-Frequency Lines from the Field of a Perfect Conductor—H. Buchholz. (*Arch. Elektrotech.*, vol. 39, pp. 79–100 and 202–215; September and December, 1948.)

621.315.221:538.541 3334
Electromagnetic Eddy-Current Fields of Spiral Form—P. Jacottet. (*Arch. Elektrotech.*, vol. 39, pp. 8–26; June, 1948.) Calculation of the effects produced in a cable sheath of finite thickness by alternating currents in a twisted pair of conductors.

621.392.26† 3335
Contributions to the Theory of Waveguides: Parts 1–4—L. Infeld, A. F. Stevenson, J. L. Synge, A. F. Stevenson, W. Z. Chien, and L. Infeld. (*Canad. Jour. Res.*, vol. 27, pp. 69–129; July, 1949.) Part 1: "Radiation from a Source inside a Perfectly Conducting Wave Guide of Rectangular Section." The field inside a semi-infinite rectangular waveguide closed at one end by a plug is determined, the current distribution in the source being regarded as known. The walls of the guide and the plug are regarded as perfectly conducting. Three different methods of solution are given. The radiation resistance is then deducted from energy considerations. In particular, an expression is derived for the radiation resistance of a linear antenna perpendicular to the wider face of the plug, fed at the point of entry. It is assumed that the antenna current is sinusoidal and that only the fundamental H-wave is transmitted by the guide.

Part 2: "A General Method for Calculating the Impedance of an Antenna in a Wave Guide of Arbitrary Cross Section." One of the methods of part 1 is extended to the case of a guide of arbitrary cross-section. The general problem of calculating radiation resistance and reactance is discussed.

Part 3: "The Resistances of Antennae of Various Shapes and Positions in Rectangular and Circular Wave Guides." Formulas are given for antennas with various assumed current distributions.

Part 4: "The Impedance of a Rectangular Wave Guide with a Thin Antenna." Contains explicit calculations for the impedance of a linear antenna in a rectangular waveguide. Appendices by J. R. Pounder and A. F. Stevenson are included.

621.392.26† 3336
The Duo-Mode Exciter—W. A. Hughes and M. M. Astrahan. (Proc. I.R.E., vol. 37, p. 1031; September, 1949.) A device is described for propagating the TE_{10} and TE_{20} modes independently in the same waveguide. The voltage S.W.R. is low over a wide frequency range and there is a little crosstalk between inputs. Performance is discussed.

621.392.26† 3337
Notes on "Wave Guides for Slow Waves"—W. Walkinshaw. (*Jour. Appl. Phys.*, vol. 20, pp. 634–635; June, 1949.) Comment on 1584 of July (Brillouin).

621.392.26†:621.3.09 3338
The Transverse Field in Waveguides of Circular Cross-Section—P. Jacottet. (*Arch. Elektrotech.*, vol. 39, pp. 108–115; September, 1948.) The similarities and differences in the behavior of the transverse fields of em oscillations of the $E_{m,n}$ and $H_{m,n}$ types are enumerated; they provide a basis for the development of field diagrams and for the clsssification of the natural oscillations according to wave type and order number. From the equation for the family of curves for the transverse field, the electric field diagram for $H_{m,n}$ waves is calculated and illustrated for the values $m=1$; 2; $n=1$; 2.

621.396.67 3339
Microwave Lenses—C. Susskind. (*Wireless World*, vol. 55, pp. 370–372; October, 1949.) A brief general survey of the three main types.

621.396.67 3340
Ground Plane Field of the Wide Angle Conical Dipole.—P. D. P. Smith. (*Jour. Appl. Phys.*, vol. 20, p. 636; June, 1949.) A formula is deduced from the work noted in 644 of April; assumptions are stated.

621.396.67 3341
Comments on Biconical Antennas—P. D. P. Smith. (*Jour. Appl. Phys.*, vol. 20, p. 633; June, 1949.) Reply to comment by Tai (1588 of July) on 644 of April.

621.396.671 3342
Impedance Transformation in Folded Dipoles—R. Guerthler. (*Jour. Brit. I.R.E.*, vol. 9, pp. 344–350; September, 1949.) The impedance of a folded dipole relative to that of a simple dipole can be adjusted by using conductors of different diameters for the separate elements of the folded dipole. Increased impedance ratios can be obtained by using additional elements. The impedance ratio can be obtained from the current ratio; formulas are derived. Practical examples are given. Reprinted from Proc. I.R.E. (Australia), April, 1949.

621.396.671 3343
A Study of the E.M.F. Method—C. T. Tai. (*Jour. Appl. Phys.*, vol. 20, pp. 717–723; July, 1949.) The method is applied to the determination of the impedance of a thin biconical antenna as first suggested by Schelkunoff (1049 of 1942). The various components of the currents flowing in the antenna are considered, and the sinusoidal part is shown to predominate. The total current at the end of the antenna, where the lateral surface of the cone and the spherical cap meet, is not identically zero but vanishes with the reciprocal of the square of the characteristic impedance of the cone. The computation of various functions involved is discussed.

621.396.671 3344
Some Aids in Sketching Field Strength Diagrams—F. Duerden. (*Electronic Eng.* (London), vol. 21, pp. 375–378; October, 1949.)

621.396.677 3345
Dielectric Aerials—H. Aberdam. (*Télév. Franç.*, no. 50, pp. 11–15; August, 1949.) A general survey of their properties, with particular reference to an unpublished paper by O. Zinke. See also 20 of February (Zinke), 1602 of July (Watson and Horton), and 1604 of July (Mallach).

621.396.679.4:621.396.931:621.396.611.4 3346
Cavity Resonators in Mobile Communications—H. Magnuski. (*See* 3378.)

CIRCUITS AND CIRCUIT ELEMENTS

621.3.016.352 3347
Stability Criterion, particularly for Control Circuits with a Prescribed Degree of Stability—A. Leonhard. (*Arch. Elektrotech.*, vol. 39, pp. 100–107; September, 1948.) Extension of previous work (567 of 1946) gives an improved stability criterion. In the characteristic equation of the system considered, the variable p is replaced by the value $\omega(j-\rho)$, where $\pi\rho$ is the logarithmic decrement, and ω is varied from 0 to ∞. The characteristic curve is thus derived simply. The behavior of this curve indicates whether all oscillations exhibit at least a prescribed attenuation, or what oscillations satisfy the required conditions and what do not. In addition, approximate values for a pair of complex roots can be derived from the curve. Practical examples illustrate the application of the method.

621.3.016.352:681.142 3348
Stabilization of Simultaneous Equation Solvers—G. A. Korn. (Proc. I.R.E., vol. 37, pp. 1000–1002; September, 1949.) In order to find out whether a number of identical amplifiers can form a stable multiple-loop feedback system for solving simultaneous equations whose coefficients form a positive definite matrix, it is only necessary to test the stability of one such amplifier with simple feedback, when this amplifier is used to solve a single simple equation.

621.314.2:629.135 3349
Small Power Transformers for Aircraft Electrical Equipments—A. L. Morris. (*Proc. IEE* (London), part II, vol. 96, pp. 413–422; June, 1949. Discussion, pp. 422–425.) The effects of climate and altitude on transformers are discussed. To obtain the smallest possible transformer for a given rating, high-grade magnetic materials should be used for the cores, and high-temperature insulating materials, such as silicone products, should replace purely organic insulating materials. A design for a 500-VA 1,600-cps transformer is included. Reprinted, *ibid.*, part III, vol. 96, pp. 279–288; July, 1949. Discussion, pp. 288–291.

621.314.3† 3350
The Theory of Magnetic Amplifiers and Some Recent Developments—E. H. Frost Smith. (*Jour. Sci. Instr.*, vol. 25, pp. 268–272; August, 1948.)

621.314.3† 3351
An Analysis of Interlinked Electric and Magnetic Networks with Application to Magnetic Amplifiers—D. W. ver Planck and M. Fishman. (Proc. I.R.E., vol. 37, pp. 1021–1027; September, 1949.) Full paper: summary noted in 2447 of October.

621.314.3† 3352
The Amplistat—A Magnetic Amplifier—R. E. Morgan. (*Elec. Eng.*, vol. 68, pp. 663–667; August, 1949.) The theory of the saturable-core reactor amplifier is briefly outlined. Several typical circuits for single-phase amplistats are given. The effects of the supply voltage and frequency, and of the load impedance, are illustrated by graphs. Amplifications as high as 10^{12} have been obtained with a single stage. Advantages include long life, no starting delay, no moving parts, ruggedness, and quiet operation.

621.316.86 3353
Thermistors: Properties and Uses of Negative-Temperature-Coefficient Resistors—(*Wireless World*, vol. 55, pp. 405–407; October, 1949.)

621.316.86 3354
Thermistors as Components Open Product Design Horizons—K. P. Dowell. (*Elec. Mfg.*, vol. 42, pp. 84–91; 212; August, 1948. Bibliography, pp. 212, 216.) A general discussion of thermistor properties and of various applications. See also 2156 of September (Butler).

621.316.86:551.508 3355
Thermistors as Instruments of Thermometry and Anemometry—W. B. Hales. (*Bull.*

Amer. Met. Soc., vol. 29, pp. 494–499; December, 1948.) For thermometry, the standard current through the Type V-560 thermistor bead is 0.02 mA which is nearly the maximum for which electrical heating of the bead can be neglected; the change in resistance with changing ambient temperature is detected by a bridge circuit. For anemometry the current varies from 3.8 mA at a temperature of 180°C for zero wind velocity to 0.5 mA at 100°C for a wind velocity of 40 m.p.h. The calibration is little affected by variations in ambient temperature.

621.318.42.011.3 3356

Approximate Formulae for Calculation of the Inductance of Circular Coils—E. Löfgren. (*Rev. Gén. Élec.*, vol. 58, pp. 305–315; August, 1949.) Examination of the approximate formulas hitherto available shows that none of them gives an accuracy within 3 per cent for all the usual forms of coil. To ensure even this accuracy at least four different formulas, depending on the type of coil, are necessary. Starting from known series for the calculation of inductance, a series for the reciprocal of the inductance is derived; this has the advantage that the first terms constitute a first approximation which can be used even outside the limits of convergence of the series. Two alternative inductance formulas with maximum error not exceeding 1 per cent are given. One gives results within 1 per cent even for coils of length or breadth much greater with respect to the mean diameter than is usual in practice. The other formula is simpler, but is limited to coils of normal form. It can, however, be written so as to be applicable to coils whose winding cross-section is not rectangular. A complementary formula for coils of small dimensions is also given, as well as a correction formula taking account of the insulation between turns.

621.318.572:539.16.08 3357

Electronic Counters for Pulses—P. Naslin and A. Peuteman. (*Onde Élec.*, vol. 29, pp. 330–335; August and September, 1949.) Continuation of 2740 of November. A detailed description is given of a counter chronometer constructed by the electromechanics section of the Laboratoire Central de l'Armement. The chronometer comprises a stabilized 100-kc oscillator from which pulses of the same frequency are derived, an electronic counter followed by a mechanical counter, and an electronic interrupter which, when closed, connects the oscillator to the counter. The counter includes a binary system of six double-triode flip-flop circuits, a decade system with two thyratron stages, and a 6-disk mechanical counter, giving a total counting time of 64,000 sec. for a pulse spacing of 10 μs. Typical measurements with the chronometer are described briefly.

621.318.572:621.396.611.4 3358

Microwave Secondary-Emission Switch—(*Electronics*, vol. 22, pp. 186, 190; October, 1949.) A cavity resonator of special shape has a gap whose faces consist of beryllium copper, which has a secondary-emission ratio of 3.5. The electron current in the gap and other operating characteristics are shown as functions of the incident power. The design is due to B. D. Steinberg.

621.319:679.5 3359

Plastic Electrets are nearing Industrial Application—T. A. Dickinson. (*Elec. Mfg.*, vol. 42, pp. 101–103; August, 1948.) Full paper; summary noted in 1307 of June.

621.392:621.3.015.3 3360

Synthesis of n-Reactance Networks for Desired Transient Response—P. R. Aigrain and E. M. Williams. (*Jour. Appl. Phys.*, vol. 20, pp. 597–600; June, 1949.) A method is given for this synthesis in the cases for which the Laplace transform of the input impulse contains either no poles or one pole at $s=0$. The relation between this method and synthesis techniques based on steady-state considerations is discussed. The method is illustrated for a simple case.

621.392:621.317.784 3361

Novel Multiplying Circuits with Application to Electronic Wattmeters—M. A. H. El-Said. (PROC. I.R.E., vol. 37, pp. 1003–1015; September, 1949.) Anode current is exponentially related to anode voltage for a diode operated in the retarding-field region. The development of the corresponding exponential mode of operation of multigrid tubes is discussed, for which anode current is accurately proportional to the product of a linear function of anode voltage and an exponential function of grid voltage over wide ranges. Various circuits using a single multigrid tube in this mode of operation are analyzed. Accurate compensation for inherent anode rectification due to the curvature of the grid-voltage characteristic is considered. Wattmeters based on this principle have predictable performance over the frequency range 20 cps to 50 Mc, absorb a very small fraction of the measured power, and can be used to measure very small powers.

621.392:621.385 3362

The Sensitivity Limit of Basic Valve Circuits—W. Kleen. (*Frequenz*), vol. 3, pp. 209–216; July, 1949.) Consideration of grounded-cathode, grounded-grid, and grounded-anode circuits shows that for given tubes and given input circuits the sensitivity limit for long waves is independent of the particular arrangement used, and nearly independent in the region of finite electron transit times. The question of the coherence or incoherence of noise currents originating from the same source is also discussed.

621.392.015.3 3363

On the Optimum Response of a Circuit with Limited Pass Band to a Heaviside Pulse—J. Laplume. (*Compt. Rend. Acad. Sci.* (Paris), vol. 229, pp. 351–352; August 1, 1949.) Formulas are derived for the output signal with minimum distortion. Calculation of the spectrum of this output signal then enables the circuit gain to be found.

621.392.4/.5 3364

The Synthesis of Passive Two-Poles by means of Networks containing Gyrators—B. D. H. Tellegen. (*Philips Res. Rep.*, vol. 4, pp. 31–37; February, 1949.) Any passive two-pole of order n may be realized by connecting a resistance across one pair of terminals of a resistanceless quadripole of the same order which may contain gyrators. The general two-pole of order n can thus be realized by one network containing the minimum number of elements, namely 1 resistor, n capacitors and inductors, and n ideal transformers and gyrators.

621.392.4 3365

Realization of Linear Two-Pole Networks with Prescribed Frequency Dependence, taking account of Losses in Coils and Capacitors—Nai-Ta Ming. (*Arch. Elektrotech.*, vol. 39, pp. 359–387; April, 1949.)

621.392.5 3366

Reactance Quadripole with Given Blocking Points and Given Unipolar No-Load or Short-Circuit Resistance—H. Piloty. (*Arch. Elek.* (Übertragung), vol. 1, pp. 59–70; July and August, 1947.) Consideration of the conditions under which such a quadripole is physically realizable.

621.392.5 3367

Realization of Linear Four-Pole Networks with Prescribed Frequency Dependence, taking account of Corresponding Losses in all Coils and Capacitors—Nai-Ta Ming. (*Arch. Elektrotech.*, vol. 39, pp. 452–471; 1949.)

621.392.5 3368

On the Design of Networks for Constant Time Delay—M. H. Hebb, C. W. Horton, and F. B. Jones. (*Jour. Appl. Phys.*, vol. 20, pp. 616–620; June, 1949.) The "group delay time" represents the delay for a signal composed of a narrow band of frequencies. The deviation of the "group delay time" from a constant value is investigated as a function of the circuit parameters for a simple LC line, an m-derived section, a type-B compensating network, and for two other networks. The relative performance of these networks and properties of the image impedance are discussed.

621.392.52 3369

Design of Dissipative Band-Pass Filters Producing Desired Exact Amplitude-Frequency Characteristics—M. Dishal. (PROC. I.R.E., vol. 37, pp. 1050–1069; September, 1949.) 1948 IRE National Convention paper. A basic method is described for obtaining the exact values required for all circuit constants in a band-pass network using n finite-Q resonant circuits, to obtain either the critical-shape-coupled type of response, or the over-coupled type. The general equation giving the gain obtained with the desired response shape is derived, together with equations for the associated phase characteristics. The particular equations for designing single-, double-, triple-, and stagger-tuned networks to produce either of the above amplitude-response shapes are tabulated. The new design method and that using the poles of the network are compared.

621.392.52 3370

L-Section Low-Pass Filter Design—P. G. Sulzer. (*Communications*, vol. 29, pp. 22–25, 32; July, 1949.) The theory of such filters is discussed, with special reference to the effect of various types of terminating impedance. Applications as power-supply filters, decoupling filters, and transmission-type filters are considered. Design charts are provided.

621.392.52 3371

Smoothing Circuits: Resistance-Capacitance—"Cathode Ray." (*Wireless World*, vol. 55, pp. 389–393; October, 1949.) A nonmathematical discussion intended to lead up to a "reliable skeleton of information on filters." The calculation of the best number of sections for such circuits is considered.

621.392.52:621.396.41 3372

Significance and Application of Frequency Filters in U.S.W. Multichannel Systems—F. Staub. (*See* 3545.)

621.392.52.012 3373

Use of a Mechanical Harmonic Synthesizer in Electric Wave Filter Analysis—S. L. Brown and J. M. Sharp. (*Jour. Appl. Phys.*, vol. 20, pp. 578–582; June, 1949.) The frequency is replaced by R-r cos θ, where R and r are arbitrary and only θ varies. The synthesizer has 15 sine and 15 cosine harmonic elements for θ. The cut-off frequency, attenuation and phase shift of a filter are determined by plotting with the synthesizer the numerator and denominator of the quantity $(x_1/-4x_2)^{\frac{1}{2}}$, where x_1 is the series reactance and x_2 the shunt reactance. The image impedances can be similarly obtained.

621.392.52.029.64 3374

Microwave Filter Theory and Design—J. Hessel, G. Goubau, and L. R. Battersby. (PROC. I.R.E., vol. 37, pp. 990–1000; September, 1949.) A theory of waveguide filters with identical links, and of matching such filters to a transmission line. "Wave matrices" are used in preference to lumped-element theory; the elements of these matrices are closely related to the reflection and transmission coefficients and can be determined more directly than the coefficients of an impedance or admittance matrix. The em state of the impedors and transducers is de-

scribed by relations between the incident and reflected waves; each transformation by a line section results only in a phase shift of these waves. Each filter stage is characterized by two angles which can be determined by simple measurements. Formulas relating to insertion properties are given in terms of these angles. The theory is applied to direct and quarter-wave coupled band-pass iris filters; good agreement with experimental results is obtained.

621.396.611:621.392.52 3375

On the Connection between Oscillators and Filters—W. Herzog. (*Arch. Elek.* (Übertragung), vol. 1, pp. 47–58; July and August, 1947.) The use of quadripole filters as oscillators is investigated and general oscillation conditions are deduced. A relation is established between these conditions and the transmission properties of the network when used as a filter. The advantages of bridge oscillators such as that of Meacham (263 of 1939) are pointed out. A circuit is considered which permits the selection of either of two frequencies of equal oscillation strength.

621.396.611.1 3376

Resonance Phenomena in Oscillatory Circuits—E. de Gruyter. (*Bull. Schweiz. Elektrotech. Ver.*, vol. 39, pp. 791–801; November 27, 1948. In German, with French summary.) Detailed treatment of (a) voltage variations at the circuit terminals and across each circuit element, (b) current and phase in the various elements, and (c) active power in the circuit. The equations for complex oscillatory circuits are derived from those for elementary circuits, suitable functions being introduced to take account of losses.

621.396.611.3 3377

Graphical Analysis of Tuned Coupled Circuits—A. E. Harrison and N. W. Mather. (Proc. I.R.E., vol. 37, pp. 1016–1020; September, 1949.) A new basis for normalizing the transfer admittance of two coupled tuned circuits permits the representation of this admittance by a single universal parabola in the complex plane. Within the limitations of the assumptions of high Q and small frequency deviations, data can be obtained from this parabola for different Q ratios, as well as the usual values of coupling and relative tuning. The method also simplifies the calculation of the input admittance of coupled circuits. Extension of the method to triple tuned circuits is possible, but the applicability of a single universal curve is lost.

621.396.611.4:621.396.679.4:621.396.931 3378

Cavity Resonators in Mobile Communications—H. Magnuski. (*Communications*, vol. 29, pp. 8–11; August, 1949.) A cavity resonator acts as a very-high-Q circuit which can be inserted between the transmitter and the antenna to decrease spurious emissions of the transmitter. It also enables several transmitters to use one antenna without interference. A cavity inserted between the receiver and the antenna provides high selectivity at rf level and rejects an unwanted signal before it reaches any receiver tube. Cavity resonators can also be used to reduce intermodulation interference

621.396.615 3379

Twin Oscillator.—T. Kirby. (*Electronics*, vol. 22, pp. 170, 182; October, 1949.) Designed primarily as a stable oscillator variable over a narrow band of frequencies. The output used is the difference between those of physically identical oscillators, one of which is padded with capacitors having zero temperature coefficient.

621.396.615:621.396.822 3380

The Influence of Thermal Resistor-Noise and of the Shot Effect on the Interference Modulation of Oscillators—A. Spälti. (*Bull. Schweiz. Elektrotech. Ver.*, vol. 39, pp. 419–427; June 26, 1948. In German, with French summary.) The magnitude of the interference modulation due to these effects is calculated and the conditions which an oscillator must satisfy for such modulation to be as small as possible are established. In practice these conditions can be fulfilled without much difficulty for frequencies up to 1,000 Mc.; at still higher frequencies AM loses its significance. For optimum results, interference modulation must be avoided not only in the transmitter oscillator, but also in the receiver heterodyne oscillator, since any modulation of the local oscillator must necessarily be transferred to the if of the receiver.

621.396.615.17 3381

A Voltage-Controlled Multivibrator—J. M. Sturtevant. (*Electronics*, vol. 22, pp. 144, 158; October, 1949.) Regenerative feedback may be applied to multivibrators and blocking oscillators to give a linear relation between frequency and the first or second power of an input dc voltage. Theory and practical circuits are discussed.

621.396.619.11.029.64 3382

Amplitude Modulation of Centimetre Waves—P. O. Hawkins and C. C. Costain. (*Nature* (London), vol. 164, p. 356; August 27, 1949.) A beam of electrons injected into a tube containing an inert gas at low pressure attenuates centimeter waves traversing the tube, provided that the electron energy is sufficient to ionize the gas. With one arrangement, maximum attenuation achieved was 50 db, with the first 15 db linearly related to applied electron current. Applications are (a) a modulator, and (b) a wide-band attenuator of low insertion loss and quick response suitable for very short pulses.

621.396.619.23 3383

Design Equations for Reactance-Tube Circuits—J. D. Young and H. M. Beck. (Proc. I.R.E., vol. 37, pp. 1078–1082; September, 1949.) Design equations are derived for several systems of reactance-tube modulation, without using the usual approximations. Expressions are obtained empirically for the total band swept; the effect of each parameter can be directly determined. The usual simplifying relations between the impedances of the feedback network were not assumed. The critical point where a given network changes from an apparent inductance to an apparent capacitance is noted.

621.396.645 3384

The Design and Limitations of D.C. Amplifiers: Parts 1 and 2—E. J. Harris and P. O. Bishop. (*Electronic Eng.* (London), vol. 21, pp. 332–335 and 355–359; September and October, 1949.) A general review with a bibliography of 38 references. Sources of random fluctuations are considered; these are on the whole more important than sources of disturbance related to external parameters. The importance of stable power-supply voltage is stressed. Design of balanced input stages, interstage coupling, and stabilization by negative feedback are discussed.

621.396.645:578.088.7 3385

Carrier-Type D.C. Amplifier for Biological Research—C. R. Maduell, Jr., and H. M. Owen. (*Electronics*, vol. 22, pp. 128, 182; October, 1949.) For continuous recording of small, slow mechanical motions. The motion is made to change the ac impedance of two coils connected in a conventional Wheatstone bridge circuit. Resulting variations in the 1,000-cps bridge output current are amplified and rectified, and actuate a recorder.

621.396.645:621.318.572 3386

Regenerative Amplifiers—Y. P. Yu. (Proc. I.R.E., vol. 37, pp. 1046–1049; September, 1949.) If a large amount of regeneration is used, the output voltage of an amplifier can be made to change abruptly from one constant value to another when the input voltage is raised to a critical value, and to change back to its original value when the input is reduced to another critical value. This principle is applied to (a) indication of the instant when two voltages become equal, (b) a peak voltmeter circuit, and (c) a pulse-width discriminator circuit.

621.396.645:621.385.029.631.64 3387

Double-Stream Amplifiers—J. R. Pierce. (Proc. I.R.E., vol. 37, pp. 980–985; September, 1949.) In the structure here analyzed, vm of the two streams occurs when they pass across the gap between the grids of the input resonator R_1 which is fed by input line L_1. This vm sets up an increasing space-charge wave, which grows in the space between input resonator R_1 and output resonator R_2. The convection current associated with the wave excites resonator R_2 and so transfers power to the output line L_2. The electron streams are collected on an anode. Formulas useful for evaluating the gain are given. See also 2486 of October (Pierce and Hebenstreit) and 2487 of October (Hollenberg.)

621.396.645:621.385.029.63/.64 3388

The Double-Stream Amplifier—A. V. Hollenberg. (*Bell. Lab. Rec.*, vol. 27, pp. 290–292; August, 1949.) See 2487 of October.

621.396.645:621.396.9 3389

Considerations in the Design of a Radar Intermediate-Frequency Amplifier—A. L. Hopper and S. E. Miller. (Proc. I.R.E., vol. 37, p. 1069; September, 1949.) Correction to 690 of 1948.

621.396.662.2 3390

The Theory and Design of Progressive and Ordinary Universal Windings—A. W. Simon. (Proc. I.R.E., vol. 37, pp. 1029–1030; September, 1949.) Comment on 1332 of 1948 (Kantor).

621.396.69]+621.317.7+621.38:061.4 3391

Electronic Equipment at Radiolympia, 1949—(*See* 3472.)

621.396.813 3392

On the Connection between Amplitude and Phase Distortion—K. W. Wagner. (*Arch. Elek.* (Übertragung), vol. 1, pp. 17–28; July and August, 1947.) The complex transfer function of an electrical transmission circuit is

$$f(j\omega) = P + jQ = ^{b+ja}$$

The system is free from distortion when db-$d\omega = 0$ and $da/d\omega =$ const. Amplitude distortion occurs if the first of these conditions is not satisfied and phase distortion when the second does not hold. P and Q, and hence b and a, are not independent of one another; f is an analytical function of the complex frequency $\zeta = \gamma + j\omega$; hence P and Q are related by the Cauchy-Riemann differential equations. The theory of functions can be used to determine either P or Q when the other is given. The calculation can also be effected by use of the Laplace transformation and Fourier integral.

The above considerations are applied to the ideal low-pass filter, which provides distortion-free transmission between $\omega = 0$ and $\omega = \Omega$ and stops all higher frequencies. It is shown that freedom from distortion can only be reached approximately; the distortion increases with increasing frequency and may be very great near the limiting frequency. The region of great distortion decreases with increasing order number k of the filter; for the ideal filter at the limiting frequency, $a = k\pi$. The degree of approximation to the ideal filter for different values of k is shown by formulas and curves. The discussion applies to any type of filter network.

621.397.645 3393

Cathode Neutralization of Video Amplifiers—J. M. Miller. (Proc. I.R.E., vol. 37, pp. 1070–1073; September, 1949.) The usual cathode bypass capacitors are replaced by a resistor

connected from each cathode to the cathode of the next stage. No gain need be sacrificed, and lf phase shift is greatly reduced. Phase shift and hf response are improved by adding a small capacitance in parallel with a portion of the intercathode resistance. Gain and stability equations are derived and a practical circuit diagram is given.

621.397.645 3394
Design of an [television] I.F. Amplifier using Stagger-Tuned Circuits—R. Aschen. (*TSF Pour Tous*, vol. 25, pp. 293–295; September, 1949.) To cover a bandwidth of 46 Mc to 56 Mc with an attenuation of 3 db and an actual gain of 60 db, four EF42 tubes, with five tuned circuits, are used. Graphs are given from which the pass band and resistance of each stage are calculated. See also 678 of 1947 (Baum).

621.397.645:621.3.015.3 3395
A New Figure of Merit for the Transient Response of Video Amplifiers—R. C. Palmer and L. Mautner. (Proc. I.R.E., vol. 37, pp. 1073–1077; September, 1949.) 1948 IRE National Convention paper. The figure of merit F proposed is $F=a\,e^{-b\gamma^2}/\tau$, where a, b are constants, τ is the rise time and γ the fractional overshoot. For a shunt-peaked stage, F is a maximum for an overshoot of about 2 per cent.

621.396.001.4 3396
Radio Servicing- Theory and Practice [Book Review]—A. Marcus, Publishers: Prentice-Hall, New York, 1948, 752 pp., $5.95. (Proc. I.R.E., vol. 37, p. 1038; September, 1949.) Intended "for those who are not beginners in radio nor yet advanced enough to study the subject on an engineering level." A clear nonmathematical discussion of tubes and their use as rectifiers, AM detectors, amplifiers, and oscillators. Power supplies, receivers, amplifiers, components, special tubes, tuning and control, servicing procedures, and repair and alignment technique are also considered.

GENERAL PHYSICS

53.081+621.3.081 3397
The Development of Electrical Units in the Last Hundred Years. The Change in the Electrical Units on 1st January 1948—U. Stille. (*Arch. Elektrotech.*, vol. 39, pp. 130–164; September, 1948.) Review and discussion, with a comprehensive conversion table from international to absolute units.

53.081+621.3.081 3398
On the Replacement of the International by the Absolute Electrical Units—H. v. Steinwehr. (*Arch. Elektrotech.*, vol. 39, pp. 27–30; June, 1948.)

53.081+621.3.081 3399
Old and New Electrical Units. A Review—J. Fischer. (*Arch. Elektrotech.*, vol. 39, pp. 340–358; February, 1949.)

534.2.001.8:621.391 3400
On the Analogy between Angle-of-Incidence and Frequency Problems—L. Cremer. (*Arch. Elek.* (Übertragung), vol. 1, pp. 28–47; July and August, 1947.)

537.122 3401
Recent Developments in the Theory of the Electron—V. F. Weisskopf. (*Rev. Mod. Phys.*, vol. 21, pp. 305–315; April, 1949.)

537.311.62 3402
Skin Effect in Conductors and in Insulators with Conducting Surface-Layer—W. Dällenbach. (*Bull. Schweiz. Elektrotech. Ver.*, vol. 39, pp. 439–446; July 10, 1948. Correction, *ibid.*, vol. 39, p. 481. In German, with French summary.) Formulas are derived for the loss per cm^2 for plane surfaces, taking account of displacement currents. In the case of a highly conductive surface layer on a material of poor or zero conductivity, the loss due to the Kelvin effect reaches a minimum value for a certain thickness of the surface layer. For greater thicknesses the loss increases again.

537.311.62 3403
Relaxation in the Anomalous Skin Effect—K. F. Niessen. (*Philips Res. Rep.*, vol. 4, pp. 38–48; February, 1949.) Comment on 1014 of 1948 (Pippard). If for infinitely long free paths of the electrons the relaxation is taken into account, a skin impedance independent of the conductivity is found. The influence of relaxation on Pippard's concept of ineffectiveness is considered.

537.312:621.315.61 3404
Electron Bombardment Conductivity—F. Ansbacher and W. Ehrenberg. (*Nature* (London), vol. 164, pp. 144–145; July 23, 1949.) Typical results obtained under dc conditoins with very thin dielectric films sandwiched between conducting layers are shown graphically and discussed.

537.523.3 3405
Point-to-Plane Corona Onsets—W. N. English and L. B. Loeb. (*Jour. Appl. Phys.*, vol. 20, pp. 707–711; July, 1949.) The effect of point material and point radius on positive and negative intermittent corona onset potentials in air at atmospheric pressure is investigated.

537.525.92 3406
Space-Charge Wave Amplification Effects—V. A. Bailey. (*Phys. Rev.*, vol. 75, pp. 1104–1105; April 1, 1949.) Discussion of the relation between the work of Haeff (1204 of May), Pierce (689 of April), and the author's theory of plane waves in an ionized gas (2785 of November). To explain solar and other noise, it is not necessary to postulate interaction of two or more different components of the stream.

537.533 3407
Thermionic Emission—C. Herring and M. H. Nichols. (*Rev. Mod. Phys.*, vol. 21, pp. 185–267; April, 1949. Bibliography, pp. 267–270.) A detailed and critical review of present knowledge of the emission from clean metals.

Chapter 1: An exposition of the thermodynamic principles underlying emission from uniform surfaces.

Chapter 2: The experimental evidence for nonuniformity of the properties of the different crystal surfaces of the same metal is summarized and the way in which this affects the interpretation of surface phenomena is indicated.

Chapter 3: The experimentally-determined emission constants for clean metals, published since 1935, are tabulated and the methods by which they were obtained are considered.

Chapter 4: Discussion of developments of modern quantum theory which have a bearing on thermionic and related phenomena.

538.11 3408
Concerning the Perfect Magnet—É. Brylinski. (*Rev. Gén. Élec.*, vol. 58, pp. 315–320; August, 1949.) The classic definition of such a magnet is recalled and consequences of the definition are examined. Numerical data for some steel magnets are discussed. A formula for the internal and external magnetic energy of the perfect magnet is given which is in agreement with the classic formula.

538.569.4.029.64:546.171.1 3409
Pressure Broadening in the Inversion Spectrum of Ammonia—H. Margenau. (*Phys. Rev.*, vol. 76, pp. 121–124; July 1, 1949.) A formula is developed to explain the low-pressure line widths observed by Bleaney and Penrose (1916 of 1948). Assumptions are those characteristic of the statistical theory of pressure broadening.

538.569.4.029.65†:546.21 3410
The Microwave Absorption Spectrum of Oxygen—M. W. P. Strandberg, C. Y. Meng, and J. G. Ingersoll. (*Phys. Rev.*, vol. 75, pp. 1524–1528; May 15, 1949.) The absorption for $\lambda \approx 5$ mm was measured. Results agree satisfactorily with the theoretical curves of Van Vleck (3098 of 1947) for an assumed line breadth $\Delta v = 0.015$ to 0.02 cm^{-1}. Observations of the absorption in mixtures of oxygen and nitrogen indicate a disparity of O_2 and N_2 collision cross-sections. Nitrogen may cause an anomalous narrowing of the oxygen absorption line.

538.6 3411
Penetration of an Alternating Magnetic Field into Solid Iron with Permeability dependent on the Field Strength—F. Nechleba. (*Arch. Elektrotech.*, vol. 39, pp. 301–318; February, 1949.)

GEOPHYSICAL AND EXTRA-TERRESTRIAL PHENOMENA

523.72:621.396.822 3412
On the Origin of Solar Radio Noise—A. V. Haeff. (*Phys. Rev.*, vol. 75, pp. 1546–1551; May 15, 1949.) Observed anomalous rf radiations from the sun are associated with sunspot activity and are believed to be generated within intermingling streams of charged particles. Such streams can greatly amplify initial space-charge fluctuations. The generation of rf energy in this way is considered theoretically; the most intense radiation is thus predicted in the frequency range 30 to 60 Mc and the corresponding absolute value of radiation intensity at the earth's surface is calculated as 7×10^{-22} to 2×10^{-22} W/cm^2 per cps. These values agree well with measurements. A formula for the most probable spectral distribution of the anomalous rf radiation is given.

523.72.029.6:621.396.822 3413
The Significance of the Observation of Intense Radio-Frequency Emission from the Sun—M. Ryle. (*Proc. Phys. Soc.*, vol. 62, pp. 483–491; August 1, 1949.) Various theories are summarized. The mechanisms proposed for the maintenance of coherent electron oscillations are examined in detail; they explain adequately the oscillations observed in discharge tubes but fail to explain satisfactorily the maintenance of electron oscillations in the solar corona. Observed results can only be accounted for by the occurrence of electron temperatures up to 10^{10} deg. K. in the solar corona near sunspots. The maintenance of such temperatures was considered in an earlier paper (696 of April).

523.746"1948" 3414
Final Relative Sunspot-Numbers for 1948—M. Waldmeier. (*Jour. Geophys. Res.*, vol. 54, pp. 187–189; June, 1949.)

523.746"1949.01/.03" 3415
Provisional Sunspot-Numbers for January to March 1949—M. Waldmeier. (*Jour. Geophys. Res.*, vol. 54, p. 192; June, 1949.)

523.75:621.396.11.029.45 3416
The Study of Solar Flares by means of Very Long Radio Waves—R. N. Bracewell and T. W. Straker. (*Mon. Not. R. Astr. Soc.*, vol. 109, pp. 28–45; 1949. Anomalies in the phase of the signal received from a very-long-wave transmitter provide a valuable indication of the occurrence of solar flares. Attempts have been made to correlate the starting times of the anomalies with the reported times of solar flares observed visually, but these have not been very fruitful. Better agreement is obtained in comparisons with photometric measurements of the width of the Hα line.

The size, duration, and time of growth of phase anomalies are analyzed statistically and it is concluded that observation of these anomalies provides the best available method of continuously monitoring solar activity.

523.78"1948.11.01":523.72.029.63 3417
Eclipse Observations of Solar Radiation at a Wave-Length of 50 cm.—W. N. Christiansen, D. E. Yabsley, and B. Y. Mills. (*Nature* (Lon-

don), vol. 164, pp. 569–570; October 1, 1949.) Discussion of 50-cm solar radiation observed at 3 Australian stations during the partial eclipse of November 1, 1948. Departures from a smooth curve in the records were used to locate small areas of greater radio brightness than their background; most of these areas were close to optical features on the sun's disk. The average temperature of the bright areas was estimated at 5×10^6 deg. K. The existence of limb brightening was neither proved nor disproved. A full account of this work will be published elsewhere.

550.38"1948.10/.12" 3418
Selected Days, Preliminary Mean *K*-Indices, and Preliminary *C*-Numbers for Fourth Quarter, 1948—H. H. Howe. (*Jour. Geophys. Res.*, vol. 54, pp. 189–191; June, 1949.)

550.38"1949.01/.03" 3419
Cheltenham [Maryland] Three-Hour-Range indices *K* for January to March, 1949—P. G. Ledig. (*Jour. Geophys. Res.*, vol. 54, p. 192; June, 1949.)

550.385:537.591 3420
Magnetic Storms and Cosmic-Ray Intensity—H. R. Sarna and O. Parkash. (*Nature* (London), vol. 164, pp. 588–589; October 1, 1949.) There is little correlation between changes in cosmic-ray intensity and those in geomagnetic intensity except during magnetic storms, when the correlation coefficient is sometimes positive and sometimes negative.

550.385"1949.01.03" 3421
Principal Magnetic Storms [January–March 1949]—(*Jour. Geoph. Res.*, vol. 54, pp. 193–195; June, 1949.)

551.510.52 3422
Charts of Dielectric Constant or Refractive Index of the Troposphere—A. W. Friend. (*Bull. Amer. Met. Soc.*, vol. 29, pp. 500–509; December, 1948.) For the conversion of radiosonde data of pressure, temperature, and water-vapor content to the dielectric constant or rf refractive index of the atmosphere.

551.510.535 3423
A Note on the Maximum Height of Reflection of a Radio Wave in a Curved Ionosphere Layer—J. M. Kelso. (*Jour. Appl. Phys.*, vol. 20, pp. 632–633; June, 1949.) A theoretical study based on Bouger's rule relating index of refraction, angle of incidence and layer height, and on an expression using Hacke's notation (3115 of 1948) in which the electron density is a parabolic function of height. For a plane ionosphere, reflections occur for all heights lying below the level of maximum ionization and above the bottom of the parabolic region. For a curved ionosphere there is an upper limit to the possible heights of reflection; this limit is below the level of maximum ionization.

551.510.535 3424
Tilts in the Ionosphere—W. Ross and E. N. Bramley. (*Nature* (London), vol. 164, pp. 355–356; August 27, 1949.) Simultaneous observations were made with two direction finders 10 km apart, on signals from a transmitter 700 km further north. Bearings taken during the day on hf signals reflected from the *F* region show fluctuations of a few degrees from the true great-circle bearing. These fluctuations have a period of 10 to 30 min., and are attributed to a tilting or wrinkling of the reflecting layer. The fluctuations were similar at both receiving stations, showing that a tilt of about 4° from the horizontal was substantizlaly uniform over a distance of 5 km. See also 2125 of 1947.

551.510.535 3425
The Ionosphere over Mid-Germany in July 1949—Dieminger. (*Fernmeldetech. Z.*, vol. 2, p. 284; September, 1949.) Continuation of 3131 of December. Abnormalities are noted.

551.510.535:550.38 3426
Theory of Lunar Effects and Midday Decrease in F2 Ion-Density at Huancayo, Peru—A. G. McNish and T. N. Gautier. (*Jour. Geophys. Res.*, vol. 54, pp. 181–185; June, 1949.) The diurnal variation of the earth's magnetic field gives rise to forced diffusion of ions in the F_2 layer. This may explain the midday decrease in F_2 critical frequency. When solar and lunar magnetic variations are in phase, the midday values of critical frequency are lower than when these variations are out of phase, as predicted theoretically.

551.510.535:550.38 3427
Seasonal Variation of World-Wide F2 Ionization for Noon and Midnight Hours—H. L. Lung. (*Jour. Geophys. Res.*, vol. 54, pp. 177–179; June, 1949.) For the noon hours there is a definite dip at the geomagnetic equator with a maximum on each side, as in the curves published by Appleton and Liang for the equinoctial months (1031 of 1948 and back references). A second maximum is also indicated at about 50° geomagnetic latitude. The midnight curves show only one maximum, which closely follows the solar declination. Midnight ionization appears to provide a better criterion of seasonal variation than noon ionization.

551.524.7+551.557 3428
Wind and Temperature Measurements up to 30 km.—F. J. Scrase. (*Nature* (London), vol. 164, p. 572; October 1, 1949.) Seven successful ascents to 30 km. have been made with large sounding-balloons. Results are compared with theoretical predictions. Further details are given in *Met. Mag.*, October, 1949.

551.594.6 3429
Recording of Atmospherics on board the Commandant Charcot—R. Bureau and M. Barré. (*Compt. Rend. Acad. Sci.* (Paris), vol. 229, pp. 525–527; September 5, 1949.) A short account of the results obtained on a frequency of 27 kc during a voyage from Brest round the Cape of Good Hope to Australia, thence to Adélie Land and back to Brest via the Suez Canal. In general, the daily variation is characterized by a maximum in the afternoon and a second maximum at night. Rapid rises or falls near sunset or sunrise are attributable to the difference of range of atmospherics during the day and the night.

Comparison of the curves with those obtained at Bagneux reveals remarkable similarities, even for the diurnal variations and for distances of the order of 2,500 to 5,000 km. The Bagneux records on 12.5 kc show the greatest similarity to those of the Commandant Charcot; in some cases the two records are almost identical. Sudden commencements of ionospheric perturbations observed on the ship and at Bagneux also show many instances of synchronism.

LOCATION AND AIDS TO NAVIGATION

621.396.9:621.396.645 3430
Considerations in the Design of a Radar Intermediate—Frequency Amplifier—A. L. Hopper and S. E. Miller. (Proc. I.R.E., vol. 37, p. 1069; September, 1949.) Correction to 690 of 1948.

621.396.93 3431
Medium-Frequency Crossed-Loop Radio Direction Finder with Instantaneous Unidirectional Visual Presentation—L. J. Giacoletto and S. Stiber. (Proc. I.R.E., vol. 37, pp. 1082–1088; September, 1949.) A crossed-loop collector system, electronic switch, single superheterodyne receiver, and synchronous rectifier are used. Design data and operating characteristics are considered, with details of the new components.

621.396.932 3432
Decca Radar—(*Wireless World*, vol. 55, p. 345; September, 1949.) Brief description of a low-priced equipment suitable for small vessels. Separate reflectors are provided for transmission and reception in the scanner unit.

621.396.933 3433
The Program for New Aids to Air Navigation—D. W. Rentzel. (Proc. I.R.E., vol. 37, pp. 1041–1042; September, 1949.) A brief survey of present systems and future plans. See also 2236 of September (Sandretto) and 3149 of December.

621.396.933 3434
Loran—M. Portier. (*Onde Élec.*, vol. 29, pp. 286–304; July, 1949.) An account of the principles, equipment, and operation of standard loran. This system is used by all aircraft of the Compagnie Air France on the North Atlantic route, whose pilots appreciate specially its rapidity and accuracy of measurement. Long-wave loran, which requires phase comparison in the receiver, is also mentioned.

MATERIALS AND SUBSIDIARY TECHNIQUES

537.228.1 3435
Mechanical Development of EDT Crystal Units—A. W. Ziegler. (*Bell Lab. Rec.*, vol. 27, pp. 245–250; July, 1949.) Details of the production and mounting of EDT crystals. These are being produced to replace quartz, which costs more, in telephone filter circuits, etc. Since EDT is soluble in water, the crystals may be sliced with a wet string, and the resonance frequency may be adjusted by wiping with a wet cloth.

538.221 3436
Coercive Field and Crystal Dimensions—F. Bertaut. (*Compt. Rend. Acad. Sci.* (Paris), vol. 229, pp. 417–419; August 8, 1949.) According to the theory of Neel (3151 and 3152 of 1947) the existence of high coercive fields in iron powders is related to the combination of three conditions which are essentially geometrical. Measurements on powders of spongy iron, prepared by reduction of ferrous formate in hydrogen at various temperatures, confirm Néel's theory. See also 2816 of November (Steinitz).

538.221:[621.318.22+621.318.32 3437
On New Ferromagnetic Materials—K. Sixtus. (*Arch. Elektrotech.*, vol. 39, pp. 260–266; December, 1948.) A review of the properties of permanent-magnet and core materials now available, including various special alloys, metal-powder products, and ferrites. See also 3447 of 1948 (Snoek), 2809 of November, and 3159 of December (Néel).

546.289:537.311.33 3438
Dependence of Resistivity of Germanium on Electric Field—R. Bray. (*Phys. Rev.*, vol. 76, pp. 152–153; July 1, 1949.) The Hall constant for *N*-type Ge in a transverse magnetic field of 4,600 gauss was found to decrease rapidly with increasing electric field strength. It was measured by the high-field pulse technique. The dependence of conductivity on temperature was studied for various transverse electric fields. Results appear to be consistent with the hypothesis that holes are injected into the *N*-type material from the positive metal contact, and are drawn into the material by the electric field. Similar results are obtained for *P*-type Ge. See also 264 of February (Bardeen and Brattain) and 1688 of July (Ryder and Shockley).

546.431.82:621.395.61/.62 3439
Barium-Titanate Ceramic as an Electromechanical Transducer—W. P. Mason. (*Bell Lab. Rec.*, vol. 27, pp. 285–289; August, 1949.) The effect of applied ac and dc fields on the

atomic structure of $BaTiO_3$ is considered. If 3–4 per cent of lead titanate is introduced into a sample of $BaTiO_3$, an isotropic ceramic is obtained which has permanent remnant polarization. This material can behave as an electromechanical transducer when cut into any shape. Advantages and possible applications are discussed.

546.482.21:535.215.1 3440
On the Photoelectric Properties of Cadmium-Sulphide Single-Crystals—J. Fassbender. (*Ann. Phys.* (Lpz.), vol. 5, pp. 33–50; June 15, 1949.)

620.197:679.5 3441
Potted Subassemblies for Subminiature Equipment—W. G. Tuller. (*Electronics*, vol. 22, pp. 104–105; September, 1949.) Design and construction procedures for potting are discussed fully. See also 442 of 1948.

621.315.5/.6 3442
New Materials and Their Engineering Significance—(*Nature* (London), vol. 165, pp. 514–517; September 24, 1949.) Summaries of three papers read at a meeting of the Engineering Section of the British Association, namely: —New Dielectric and Semi-conducting Materials, by R. W. Sillars; Magnetic Materials for Electrical Power Plants, by F. Brailsford; and Metals for High Duty, by R. W. Bailey.

621.315.5/.6 3443
Properties of Conductive Plastics—(*Electronics*, vol. 22, pp. 96–99; October, 1949.) Description of Markite materials, which can be moulded but have electrical resistivities comparable with that of Hg. Many of these materials can be soldered or electroplated directly. Applications include shielding, moulded circuits, resistors, and commutators.

621.315.59:621.385.032.216 3444
Semi-Conducting Properties in Oxide Cathodes—N. B. Hannay, D. MacNair, and A. H. White. (*Jour. Appl. Phys.*, vol. 20, pp. 669–681; July, 1949.) The electrical conductivity of (Ba, Sr)O has been studied as a function of temperature before and after activation with methane. (Ba, Sr)O appears to be a "reduction" semiconductor whose conduction electrons probably arise from a stoichiometric excess of (Ba, Sr) atoms in solid solution. The electrical conductivity and the thermionic emission of a (Ba, Sr)O cathode are directly proportional through three orders of magnitude of activation, as predicted by semiconductor theory.

621.315.592† 3445
Theory of Electronic Semiconductors and of their Complex Derivatives—S. Teszner. (*Bull. Soc. Franç. Elec.*, vol. 9, pp. 401–432; August, 1949.) Various theories concerning the properties of semiconductors are reviewed. The physical mechanism of the phenomena in the boundary layer of a semiconductor and in complex derivatives such as thermistors is discussed. A tentative theory is proposed which is in better agreement with experiment than many previous theories.

621.315.611.015.5 3446
The Electrical Breakdown of Solid Insulators—K. W. Wagner. (*Arch. Elektrotech.*, vol. 39, pp. 215–233; December, 1948.) General discussion, including theory of heat breakdown, effect of electric field, time effects, breakdown under alternating voltages, and influence of frequency and temperature. Curves for typical materials illustrate the effects observed.

621.315.612 3447
Review of New Developments in H. F. Ceramics—C. Schreck. (*Fernmeldetech. Z.*, vol. 2, pp. 285–295; September, 1949. Bibliography, p. 296). Discussion of the mechanical and electrical properties of materials produced in Germany, America, Russia, and England for both high-power and low-power equipment. In the case of capacitor materials the temperature coefficient of the dielectric constant is considered with particular reference to the design of temperature-compensated circuits.

621.315.614 3448
Dependence of the D.C. Resistance and Loss Angle of Paper on its Dryness and Temperature—H. Veith. (*Frequenz*, vol. 3, pp. 165–173 and 216–223; June and July, 1949.) Apparatus for measurements at temperatures between −50° and +50° C and at frequencies between 400 cps and 100 kc is described. Results are shown in numerous diagrams. The dc resistance falls exponentially with the moisture content; it decreases to about a tenth for a 1.5 per cent increase of moisture content. The temperature coefficient of the dc resistance is about the same for different moisture contents. The variation of ac resistance with humidity is much less than that for dc, particularly for low temperatures and high frequencies. The loss angle at first increases only slightly with moisture content, but beyond a certain point, which at low temperatures moves towards higher humidity values, the increase becomes somewhat greater. At low temperatures even a water content of 10 per cent produces only a slight increase of loss angle. The results are discussed with reference to the structure of cellulose.

621.315.616 3449
A Room-Temperature Transition in Polytetrafluoroethylene—H. A. Rigby and C. W. Bunn. (*Nature* (London), vol. 164, p. 583; October 1, 1949.) As the temperature increases through 20°C, the specific volume of polytetrafluoroethylene increases suddenly by about 1 per cent. X-ray-diffraction evidence that this is associated with a change in crystal structure is discussed.

621.315.616:620.197 3450
"Araldite"—C. J. Moss. (*Electronic Eng.* (London), vol. 21, pp. 389–392; October, 1949.) A synthetic resin of complex composition which does not give off water or volatile products when it sets, does not shrink much when it hardens, and adheres to metals, mica, quartz, porcelain, glass, etc. It appears to be capable of meeting all the sealing requirements for Service equipment. The material was developed by Messrs. Ciba of Basle.

621.315.616:679.5:621.791 3451
Radio Frequency Welding of Plastics—L. Grinstead and H. P. Zade. (*Jour. Brit. I.R.E.*, vol. 9, pp. 322–338; September, 1949.) Heated-tool and hot-gas welding methods depend upon the thermal properties of the material to be welded, whereas rf methods make use of the dielectric losses occurring when the material is subjected to hf fields. Rf methods are unsuitable for low-loss materials, as the rate of heating depends upon the loss factor. The effects of variation in physical constants with changing temperature and of thermal conduction of the welding electrodes on the type of rf generator required are discussed. Generator design and typical oscillator circuits are considered.

621.318.2 3452
Permanent Magnets: Properties, Design Materials—J. Fischer. (*Arch. Elektrotech.*, vol. 39, pp. 327–340; February, 1949.)

621.318.22 3453
Anisotropic Permanent Magnet Alloy—K. Hoselitz and M. McCaig. (*Nature* (London), vol. 164, pp. 581–582; October 1, 1949.) Discussion of experimental determinations of some magnetic characteristics of Alcomax which confirm and extend the conclusions of Jellinghaus (*Z. Für Metallkunde*, vol. 39, p. 52; 1948) about domain magnetization, and explain the remarkable magnetic hardness of this type of alloy.

621.383.4 3454
Theory of Photoconductivity of Layers of Semiconducting Substances—Schwarz. (*See* 3578).

621.775.7:538.213 3455
Magnetic Properties of Iron Compacts in relation to Sintering Temperature—R. Steinitz. (*Jour. Appl. Phys.*, vol. 20, pp. 712–714; July, 1949.) A paper presented at the International Powder Metallurgy conference noted in 753 of April. The relation between permeability and density for the higher sintering temperatures is discussed.

MATHEMATICS

517.93:531 3456
A Method of Equivalent Linearization for Non-Linear Oscillatory Systems with Large Non-Linearity—C. A. Ludeke. (*Jour. Appl. Phys.*, vol. 20, pp. 694–699; July, 1949.) The method described is based partly on existing nonlinear theory and partly on experimental results. See also 156 of 1948 (Minorsky).

681.142 3457
More Differential Analyzer Applications—A. C. Cook and F. J. Maginniss. (*Gen. Elec. Rev.*, vol. 52, pp. 14–20; August, 1949.) Continuation of 3114 of 1945 (Maginniss).

681.142 3458
Electronic Calculating-Machine Development in Cambridge—M. V. Wilkes. (*Nature* (London), vol. 164, pp. 557–558; October 1, 1949.) The operation of EDSAC is explained with reference to the calculation of Airy's integral $Ai(-x)$. The machine could calculate 100 values in about 4 minutes. See also 2823 of November and 3459 below.

681.142 3459
Progress in High-Speed [digital] Calculating Machine Design—M. V. Wilkes. (*Nature* (London), vol. 164, pp. 341–343; August 27, 1949.) Report of a conference held at Cambridge to mark the completion of the EDSAC. See also 3448 of 1948 (Wilkes and Renwick) and 3458 above.

681.142:518.61 3460
Precise Solution of Partial Differential Equations by Resistance Networks—G. Liebmann. (*Nature* (London), vol. 164, pp. 149–150; July 23, 1949.) The accuracy of electrical analogue methods of solving equations, normally within about 1 per cent–5 per cent, may be greatly increased by using a process of successive approximation. Solutions of some equations have been obtained within 1 or 2 parts in 10^4.

681.142:621.3.016.352 3461
Stabilization of Simultaneous Equation Solvers—Korn. (*See* 3348.)

517.392(083.74) 3462
Tables of Generalized Sine- and Cosine-Integral Functions [Book Review]—Publishers: Harvard University Press, Cambridge, Mass., 1949, 2 vols, 462 and 560 pp., $20.00. (PROC. I.R.E., vol. 37, p. 1034; September, 1949.) Tables computed to 6 decimal places on the Automatic Sequence Controlled Calculator.

MEASUREMENTS AND TEST GEAR

620.178.3:621.396.69 3463
Theory of Vibration Testing—O. Heymann. (*Frequenz*, vol. 3, pp. 196–208; July, 1949.) A general solution is obtained for the motion of elastically supported particles with one degree of freedom. Various types of applied accelerating force are considered and their effects are

discussed. Of all possible vibrations to which apparatus may be subjected, the periodic type is the most dangerous, particularly if the vibration period approaches the natural period of the apparatus in question. In this case, as in most other cases, the impulse given to the apparatus is a better index of strain than the actual acceleration. Suggestions for improved vibration tests are made.

621.317.321†:621.314.2 3464
Method for the Direct Measurement of the Active and the Reactive Voltage Drop in Transformers—F. Koppelmann. (*Arch. Elektrotech.*, vol. 39, pp. 164–183; September, 1948.) With the help of a moving-coil instrument preceded by a mechanical rectifier, both the resistive and inductive voltage drop of a transformer can be measured, not only on short-circuit but also under normal conditions of operation. The application of the method to a single-phase and to a 3-phase transformer is explained.

621.317.335.2†+621.317.374]:621.319.45 3465
Measurement of the Apparent Capacitance and the Loss Angle of Electrolytic Capacitors by the Three-Ammeter Method—V. Aschoff. (*Arch. Elektrotech.*, vol. 39, pp. 414–419; April, 1949.) A variable impedance is connected in parallel with the capacitor and adjusted till the currents in the two branches, due to an applied alternating voltage, are equal. The loss angle is then easily determined from the ratio of the main circuit current to that in either branch; simple formulas give the value and phase of the apparent resistance of the capacitor and also its apparent capacitance. Correction formulas are necessary if the meter resistances cannot be regarded as negligible.

621.317.335.3†:621.315.612 3466
Method of Determining the Dielectric Constant and Power Factor of Ceramics at 100 Mc. as a Function of Temperature—H. J. Evans. (*Jour. Amer. Ceram. Soc.*, vol. 32, pp. 262–266; August 1, 1949.) A resonance method, using a specially constructed *Q*-meter. Temperature is varied by enclosing the whole equipment in an electric heater and *Q* is measured by the frequency-variation method. Typical results are shown graphically and discussed.

621.317.4 3467
Magnetic Testing Symposium features the 1948 Meeting—(*Elec. Mfg.*, vol. 42, pp. 121–125, 220; August, 1948.) Summaries of 9 papers presented at the meeting.

621.317.44 3468
New Coil Systems for the Production of Uniform Magnetic Fields—J. R. Barker. (*Jour. Sci. Instr.*, vol. 26, pp. 273–275; August, 1949.) Systems of three or four identical air-cored circular coils are described. The required separations of these coils and the currents in them are calculated with due allowance for the finite cross-section of the coils.

621.317.44:552.12 3469
A New High Sensitivity Remanent Magnetometer—E. A. Johnson, T. Murphy, and P. F. Michelsen. (*Rev. Sci. Instr.*, vol. 20, pp. 429–434; June, 1949.) The instrument, constructed for use in a mobile field laboratory, is capable of measuring magnetic moments as small as 2×10^{-8} c.g.s. units per cm³ in rock samples of volume 25–50 cm³. Circuit and practical details are given.

621.317.44:621.318.24 3470
A Sensitive Balance for Stability Tests on Permanent Magnets—S. F. Knight. (*Proc. IEE* (London), Part II, vol. 96, pp. 635–640; August, 1949.) The force acting on a conductor in the field of the magnet is balanced against the weight of a fixed mass. An optical system is used to detect unbalance. Errors and their elimination are discussed. Typical test results indicate an overall maximum error of +0.03 per cent.

621.317.66:621.317.794.029.64 3471
Determination of Efficiency of Microwave Bolometer Mounts from Impedance Data—D. M. Kerns. (*Bur. Stand. Jour. Res.*, vol. 42, pp. 579–585; June, 1949.) The bolometer mount is regarded as a transducer which can be represented as a 2-terminal-pair network, and its parameters are determined from observation of input impedance as a function of bolometer resistance. Theory of the method is discussed and formulas are given for calculating the power-transfer efficiency from such impedance data.

621.317.7+621.38+621.396.69]:061.4 3472
Electronic Equipment at Radiolympia, 1949—(*Electronic Eng.* (London), vol. 21, pp. 367–374; October, 1949.) Brief descriptions of various exhibits.

621.317.71 3473
A Precision Automatic Electrometer—N. T. Seaton. (*Rev. Sci. Instr.*, vol. 20, pp. 500–503; July, 1949.) For measuring currents of the order 10^{-11}–10^{-11} A in terms of the time required to charge a capacitor to a certain voltage. Accuracy is within 0.1 per cent when the input FP-54 pliotron is operated under special conditions; negative feedback is used.

621.317.72:621.314.63 3474
Rectifier-Voltmeters—O. Macek. (*Frequenz* vol. 3, pp. 223–226; July, 1949.) Description of a periodic and tuned voltage and field-strength meters for dm and cm waves, using Ge rectifiers. With a resistance of 100 kΩ in series with the indicating meter, the calibration curve plotted on logarithmic paper is linear above 0.3 V. Rectifier sensitivity data for frequencies of 10 Mc and 7,000 Mc are tabulated.

621.317.725 3475
New Type of High-Voltage Voltmeter for Absolute Measurements—B. Gänger. (*Arch. Elektrotech.*, vol. 39, pp. 443–452; 1949.) A plane circular electrode, held centrally in a gap in the face of a much larger guard-ring electrode by three insulated radial springs, moves outward slightly when a third electrode of opposite polarity is brought near. The motion causes a reduction of the capacitance of a parallel-plate capacitor connected mechanically to the électrode. This capacitance change is a measure of the applied voltage and causes variation of the frequency of a hf oscillator. Theory of the method is given and the construction and calibration of practical equipment is described.

621.317.725 3476
Reflex Valve Voltmeter—M. G. Scroggie. (*Wireless World*, vol. 55, pp. 401–404; October, 1949.) An effective method of stabilization involves the addition of only one extra resistance to the simplest form of such a voltmeter. Operation for symmetrical square-wave input is described and calibration curves are given. A circuit diagram for such a voltmeter with ranges of 5 V, 20 V, and 50 V is included.

621.317.725 3477
Arrangement for Indicating Small Direct Voltages with a Pen Recorder—W. Geyger. (*Arch. Elek.* (Übertragung), vol. 3, pp. 165–173; August, 1949.) The main features of other devices of the pen-recorder type hitherto available are reviewed. The essential components of the instrument described are a magnetic amplifier, a tube amplifier and a phase-sensitive rectifier whose output is applied to a moving-coil recorder and a compensating resistor. Voltages of the order of 0.1–1.0 mV can be measured to within 1 per cent.

621.317.725:621.316.722.4 3478
Vacuum Capacitor Voltage Dividers—E. F. Kiernan. (*Electronics*, vol. 22, pp. 140, 146; September, 1949.) The ac range of a tube voltmeter can be extended and the input impedance can be simultaneously increased by using compact vacuum capacitors which can be assembled to form voltage dividers consisting of two sections in series.

621.317.734 3479
Valve Megohmmeter—W. H. Cazaly. (*Wireless World*, vol. 55, pp. 326–328; September, 1949.) A true potentiometer circuit is used with a high-impedance tube-operated indicator. Accuracy is within about 2 per cent for the range 5kΩ–5MΩ; readings are independent of battery voltage within wide limits. See also 153 of February (Spratt).

621.317.75 3480
A Direct Reading Pulse Length Meter and Shape Analyzer—R. Rudin. (*Rev. Sci. Instr.*, vol. 20, pp. 467–471; July, 1949.) An instrument designed to measure the duration of repeated current or voltage pulses, and to determine the form of the pulse by measuring the width at different heights. Comparison with other methods shows good agreement of the results.

621.317.761:513.618.1 3481
Frequency Comparison with Cycloids—W. Bader. (*Arch. Elektrotech.*, vol. 39, pp. 115–129; September, 1949.) For the comparison of two frequencies within a wide range, epicycloids or hypocycloids are generated on the screen of a cro. Comparatively large frequency ratios, either integral or fractional, can be identified quite easily. The direction of apparent rotation of the diagram seen on the screen, when the frequency ratio is near that which gives a stationary pattern, is an indication as to whether one of the frequencies is too high or too low. Complete theory of the method is given and numerous illustrations of actual patterns are included.

621.317.763 3482
Citizens Radio Wavemeter—W. B. Lurie. (*Electronics*, vol. 22, pp. 88–91; September, 1949.) Two coaxial-type wavemeters are described. Formulas are given from which the impedance can be calculated, and the effect of distortion near the end plate is considered. Accuracy with typical component values is discussed; it may easily be made adequate for use with class-B equipment. Sufficient accuracy for use with class-A transmitters can be obtained with greater construction precision. Increased sensitivity can be obtained if a fixed capacitor is used to provide some of the loading capacitance. For other articles on Citizens' Radio see 517 of March (Samuelson), 1006 of May (Hollis), and back references.

621.317.772.029.3/.4 3483
Measuring Phase at Audio and Ultrasonic Frequencies—E. R. Kretzmer. (*Electronics*, vol. 22, pp. 114–118; October, 1949.) The periodic signals to be compared are converted to square waves whose edges coincide with the points where the amplitudes of the original signals are zero. These square waves are used to form pulse trains which successively trigger a flip-flop circuit. The phase difference is read directly; it depends upon the average current through one side of the flip-flop circuit.

621.317.78:536.532 3484
Dynamic Impedance and Sensitivity of Radiation Thermocouples—P. B. Fellgett. (*Proc. Phys. Soc.*, vol. 62, pp. 351–359; June 1, 1949.)

621.317.784:621.392 3485
Novel Multiplying Circuits with Application to Electronic Wattmeters—El-Said. (*See* 3361.)

621.317.79:621.385.032.216 3486
A New Emission Microscope for Oxide Cathodes—L. Jacob. (*Jour. Sci. Instr.*, vol. 26, pp. 262–266; August, 1949.) For investigating the spatial variation of emission over an oxide surface. The action of the microscope depends on the fact that the modulator of a conventional electron gun behaves like an iris diaphragm when its potential is varied. The modulator thus controls the area of emission on the cathode surface, and enables the variations which occur over small areas to be investigated. Effects at both high and low modulations are examined for an extensive oxide surface. The instability associated with temperature-limited emission is investigated for all points on the modulation characteristic.

621.317.79:621.396.615.12 3487
Audio Signal Generator: Parts 1 and 2—M. G. Scroggie. (*Wireless World*, vol. 55, pp. 294–297 and 331–334; August and September, 1949.) Details of an af oscillator for general laboratory work, and discussion of auxiliary circuits and frequency calibration.

621.317.79:621.397.7 3488
A TV Monitor Receiver—F. C. Grace. (*Communications*, vol. 29, pp. 10–13, 34; February, 1949.) A low-gain receiver to enable the transmitter operator to assess the quality of the picture being broadcast. Requirements additional to those of good home receivers are protection against overloading by strong signals and high discrimination against unwanted rf signals.

621.385.001.4 3489
Modern Test Methods revealing the Characteristics and Limitations of V.H.F. and U.H.F. Valves—R. Remillon. (*Onde Élec.*, vol. 29, pp. 273–285 and 336–346; July, August, and September, 1949.) A review of the tests which should be carried out by tube manufacturers to give users the necessary information for predicting the operation of any particular tube under given conditions, and for indicating the limitations of the use of normal tubes at vhf. The following classes of tubes are considered: (a) receiving or low-power transmitting tubes, miniature and acorn tubes; (b) vhf transmitting triodes, tetrodes, or pentodes used in circuits with localized constants or with Lecher transmission lines; (c) vhf and uhf transmitting triodes and tetrodes used with cavity-resonator circuits; (d) klystrons and traveling-wave tubes; and (e) magnetrons.

621.396.933.001.4 3490
The Maintenance of Aircraft Radio Equipment—(*Engineering* (London), vol. 168, p. 164; August 12, 1949.) Brief description of a Marconi test set comprising a wide-range signal generator, a beat-frequency oscillator, af power meter and apparatus including a loop and a vertical antenna for testing df equipment. The set is intended primarily for use at main service depots where an adequate supply of spare units is available. For transmitter testing, additional modulator, power, and loading units are required.

531.761 3491
Electronic Time Measurements [Book Review]—B. Chance, R. I. Hulsizer, E. F. MacNichol, Jr., and F. C. Williams (Eds). Publishers: McGraw-Hill, New York and London, 1949, 538 pp., 42s. (*Nature* (London), vol. 164, pp. 594–595; October 8, 1949.) Vol. 20 of the M.I.T. Radiation Laboratory series. The book is really a sequel to that noted in 2006 of August, and "describes some very ingenious circuit developments originated both in Great Britain and in the United States." Subjects covered include electronic integrators, accurate time measurements connected with the navigation aids Gee and Oboe, automatic following of radar responses, methods of relaying an air borne p.p.i. picture to a ground station, etc.

621.317.2 3492
Radio Laboratory Handbook [Book Review]—M. G. Scroggie. Publishers: Iliffe and Sons, London, 4th edn 1948, 424 pp. (Proc. I.R.E., vol. 37, p. 1036; September, 1949.) Intended to guide a laboratory worker in setting up and properly using a laboratory. Commercially available British laboratory instruments and a few special instruments are described, and general methods of measurement considered.

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

531.717.1:539.16.08 3493
Radioactive Thickness Gage for Moving Materials—J. R. Carlin. (*Electronics*, vol. 22, pp. 110–113; October, 1949.) Material moving between an ionization chamber and a radioactive source absorbs β-particles or γ-rays in proportion to its thickness. A meter in the electrometer amplifier circuit can be calibrated to read thickness directly.

535.61-15 3494
A New Industrial Infra-Red Spectrometer—R. R. Gordon, H. Powell, and R. A. C. Isbell. (*Jour. Sci. Instr.*, vol. 25, pp. 277–282; August, 1948.) A detailed description. Results obtained with two instruments of this design on three hydrocarbons are compared with those obtained with earlier instruments, and calibration problems are considered.

539.16.08 3495
Characteristics of Halogen Counters—S. H. Liebson. (*Rev. Sci. Instr.*, vol. 20, pp. 483–484; July, 1949.)

539.16.08 3496
A Demountable Geiger-Muller Counter using Filling Gases at Atmospheric Pressure—J. F. Tait and G. H. Haggis. (*Jour. Sci. Instr.*, vol. 26, pp. 269–271; August, 1949.) For the assay of tracers emitting low-energy β-rays. A helium/ether mixture flows through the counter.

539.16.08 3497
A Circuit for the Study of the Operation of the Geiger-Muller Counter—M. A. Guimarães and P. A. Sampaio. (*Rev. Sci. Instr.*, vol. 20, pp. 485–488; July, 1949.)

539.16.08 3498
Portable Geiger Counter for Drill Holes—A. Roberts. (*Radio and Telev. News, Radio-Electronic Eng. Supplement*, vol. 13, pp. 16–17, 28; September, 1949.) For detecting radioactivity at depths down to 1,000 ft. The counting apparatus is fixed on the side of a cable drum; a single-ended self-quenching G-M tube, without preamplifier, is attached to the free end of the cable. Circuit and component details are given.

539.16.08:621.318.572 3499
Electronic Counters for Pulses—Naslin and Peuteman. (*See* 3357.)

551.508:621.316.86 3500
Thermistors as Instruments of Thermometry and Anemometry—Hales. (*See* 3355.)

551.508.1:621.396.9 3501
A Dutch Radiosonde—J. L. van Soest. (*Tijdschr. ned. Radiogenoot.*, vol. 8, pp. 305–313; October, 1940. In Dutch.) Description of apparatus for transmitting to a ground station information concerning air pressure, temperature, and humidity. A rotary switch, operated by a small $\frac{3}{4}$-W motor, provides a system of coded signals on a carrier frequency of 50 Mc.

620. 179.14:625.17 3502
Railroad Track Inspection Car—R. D Walker, Jr. (*Electronics*, vol. 22, pp. 66–68; October, 1949.) The car is self-propelled and has a special generator which passes large currents at low voltage through the rails. Any flaw, such as a transverse crack produces a characteristic variation of the flux through search coils at the mid-point of the equipment. Such variations are used to provide an indication of the location of the flaw on the tape of a recorder. Inspection is carried out at about 12 m.p.h. A sensitive hand set is provided for detailed examination of suspected flaws.

621.316.7.078 3503
Planning for Automatic Process Control—(*Electronics*, vol. 22, pp. 72–79; October, 1949.) Discussion of the general principles of various types of control devices and their relative merits typical sensing elements, automatic controllers, and correcting devices. For various applications in heavy and light industries, brief details are given of the type of equipment used; illustrative diagrams of typical installations are included.

621.317.39:531.391:531.15 3504
Rotor Balancing with an Electronic Capacitor Gauge—H. D. Warshaw. (*Rev. Sci. Instr.*, vol. 20, pp. 474–476; July, 1949.) A variable air-gap capacitance exists between an unbalanced rotating member and a stationary plate. Variations of the air-gap are used to modulate the output of an oscillator; this output is displayed on a cro. Cyclic displacements of 0.0005 in. at the circumference of the rotor are easily detected, and can be located by means of a contactor on the rotor shaft.

621.365.54† 3505
The Heating of Materials of Low Conductivity by H. F. Induction Currents—G. Ribaud. (*Jour. Phys. Radium*, vol. 8, pp. 97–101; April, 1947.) Graphs are given from which the efficiency of heating can be calculated for various substances. With frequencies of the order of 1 Mc, excellent efficiencies can be obtained with materials of resistivity from 10^7 to 10^{10} c.g.s. units and of dimensions exceeding 6 cm.

621.38.001.8 3506
Institution of Electronics (N. W. Branch). Fourth Annual Exhibition—H. Steeple. (*Nature* (London), vol. 164, pp. 525–526; September 24, 1949.) Brief descriptions of various exhibits. See also 3219 of December.

621.38.001.8:621.9 3507
Electronic Control of Machine Tools—S. A. Ghalib. (*Engineering* (London), vol. 168, pp. 173–174; August 19, 1949.) Long summary of paper presented at a joint meeting of the British Institution of Radio Engineers and the Institution of Production Engineers. Electronic applications for machine tools are mainly to (a) variable-speed drives, (b) feed control, and (c) automatic machining of a given profile. See also 1449 of June.

621.384.611.2† 3508
A Resonance Effect in the Synchrotron—E. D. Courant. (*Jour. Appl. Phys.*, vol. 20, pp. 611–616; June, 1949.) Resonance effects occurring when $n = d(\log H)/d(\log r) = \frac{3}{4}$ will reduce the intensity of the beam if the first Fourier component of the azimuth variation of n exceeds about 10^{-3}, unless the equilibrium value of n is sufficiently far from $\frac{3}{4}$.

621.384.611.2† 3509
Three Dimensional Design of Synchrotron Pole-Faces—C. Robinson. (*Proc. Phys. Soc.* (London), vol. 62, pp. 592–597; September 1,

1949.) An analytical method is outlined which takes full account of the cylindrical symmetry. Relaxation methods are used (a) to determine the exact size and shape of the lips to correct for fringing, and (b) to check the characteristics of the magnetic field. Practical examples are included.

621.385.833 3510
Field Asymmetry due to the Voltage Feeders for Electrostatic Lenses—F. Bertein. (*Compt. Rend. Acad. Sci.* (Paris), vol. 229, pp. 291–293; July 25, 1949.)

621.385.833 3511
The Illuminating System of the Electron Microscope—J. Hillier and S. G. Ellis. (*Jour. Appl. Phys.*, vol. 20, pp. 700–706; July, 1949.)

621.385.833 3512
The Scanning Principle in Ultra-Microscopy—H. Mahl. (*Elektron Wiss. Tech.*, vol. 3, pp. 350–354; September, 1949.) Short description of three types of microscope using secondary emission. The first, due to Knoll (1157 of 1936), gives an image directly on the screen of a cr tube. In the second, due to v. Ardenne (1167 of 1939), the image is recorded on a photographic film, while that of Zworykin, Hillier, and Snyder (139 of 1943) makes use of a facsimile printer.

621.385.833:061.3 3513
Conference on Electron Microscopy, Delft—V. E. Cosslett. (*Nature* (London), vol. 164, pp. 481–483; September 17, 1949.) A general account of the proceedings, with abstracts of selected papers.

621.396.9:674 3514
Metal Detector for the Lumber Industry—C. R. Schafer. (*Electronics*, vol. 22, pp. 100–103; September, 1949.) A 4-coil arrangement designed to provide more uniform sensitivity to objects embedded in logs at various angles. Complete circuit diagrams with component values and coil winding data are included. See also 3212 of 1948 (Hacks).

621.398:623.41 3515
Study and Realization of Electronic Telecontrol of Artillery—R. Aubry, G. Lehmann, and H. le Boiteux. (*Onde Elec.*, vol. 29, pp. 311–329; August and September, 1949.) Part 1: "History of the Subject and Statement of the Problem." The conditions which must be satisfied by equipment for naval vessels and the reasons which led the naval authorities to investigate electronic methods are considered. An outline of the development of suitable equipment is given.

Part 2: "Principles and Methods which have served as a Basis for Study." The theory of servomechanisms is applied to the determination of the cut-off frequency for a given precision of response and given conditions of stability. The essential characteristics of the output stage and the effects of filter inductance and motor inertia are considered and also the design of correcting networks.

Part 3: "Description of Equipment. Results Obtained." Details are given of the methods and apparatus finally adopted for gun control. Records show that considerably higher accuracy of setting was achieved than was originally specified; and the stability of the system was excellent.

621.38.001.8 3516
Electronics and their Application in Industry and Research [Book Review]—B. Lovell (Ed.). Publishers: Pilot Press, 1947; now issued by Chapman and Hall, London, 660 pp., 42s. (*Electronic Eng.* (London), vol. 21, pp. 396–398; October, 1949.) ". . . a long, soundly written, soundly edited account of recent developments, written by leading specialists. Much of it appears for the first time in book form, including many useful practical data. . . . This book can be strongly commended to all electronic engineers and to others who would gain a wider view of the possibilities of electronics, as it combines to an uncommonly high degree the virtues of scientific authority and compact, interesting presentation." Each chapter is self-contained and has an up-to-date bibliography.

PROPAGATION OF WAVES

538.566.2:512.9 3517
Tensor Field Equations in a Region of Variable Refractive Index—P. D. P. Smith. (*Jour. Appl. Phys.*, vol. 20, p. 633; June, 1949.)

538.566.3:551.510.535 3518
Propagation of Electromagnetic Waves in a Layered Medium under the influence of a Magnetic Field, for Oblique Incidence—K. Försterling. (*Arch. Elek.* (Übertragung), vol. 3, pp. 115–120; July, 1949.) The differential equations for the propagation of em waves in the ionosphere are developed and their solution, as affected by the earth's magnetic field, is discussed.

538.566.3:551.510.535 3519
Some Remarks on the Ionospheric Double Refraction: Part 1—H. Bremmer. (*Philips Res. Rep.*, vol. 4, pp. 1–19; February, 1949.) General theory is discussed, with special reference to sw propagation. Snell's law determining the normal direction of propagation is given explicitly; the connection between this direction and that of the corresponding ray is derived (a) from the mathematical theory of the characteristic surfaces of a partial differential equation, (b) by considering Fresnel's indicial surface, (c) from Fermat's principle, and (d) from the Poynting vector. For a given primary ray, the splitting into an ordinary and an extraordinary ray on reaching the ionosphere and the state of polarization on leaving it are investigated. The corresponding theory for long waves is summarized and illustrated by a numerical example.

621.396.11:551.510.52 3520
Tropospheric Measurements—P. Harbury. (*Electronics*, vol. 22, pp. 126, 128; October, 1949.) Description of a 106-Mc 100-kW pulse transmitter using a parabolic reflector 62 ft. in diameter which radiates vertically, and the associated recording system. The apparatus will be used to continue and extend the work of Friend noted in 2593 of 1948 and 1760 of July.

621.396.11:551.510.535 3521
Ionospheric Absorption—R. N. Bracewell and K. Weekes. (*Jour. Appl. Phys.*, vol. 20, p. 724; July, 1949.) Criticism of 2027 of August (McCracken).

621.396.11:551.510.535:061.3 3522
The Ionosphere and the Propagation of Radio Waves—J. A. Ratcliffe. (*Nature* (London), vol. 164, pp. 511–513; September 24, 1949.) Report of a Summer Meeting of the Physical Society at Cambridge, held to survey and co-ordinate research work in progress.

621.396.11.029.58 3523
Investigations of High-Frequency Echoes: Part 2—H. A. Hess. (PROC. I.R.E., vol. 37, pp. 986–989; September, 1949.) Continuation of 3495 of 1948. Records of telegraphy signals at frequencies between 10 and 20 Mc which show periodic variations of the field strength are investigated. See also 2608 of October.

621.396.11.029.62:621.396.1 3524
The Ad Hoc Committee Report—R. Lewis. (*Communications*, vol. 29, pp. 6–9, 30; July, 1949.) Report of a study of propagation problems for FM and television transmissions at frequencies between 50 and 250 Mc, providing data for frequency channel allocation.

Four main objectives are considered: (a) prediction of service field intensities, taking into account standard ground-wave signal range curves, effective antenna heights, and correction factors based on experimental results, (b) evaluation of random terrain variation, (c) tropospheric propagation curve evaluation, and (d) the general principles involved in combining the effects of variations of the desired signal and one or more interfering signals. Further investigation of (d) is required.

621.396.81:621.397.5 3525
Technique for TV Field Surveys—J. F. Dreyer, Jr. (*Electronics*, vol. 22, pp. 82–85; October, 1949.) Circumferential as well as radial measurements should be taken, and antennas at heights of both 10 ft. and 30 ft. should be available. In hilly country or built-up areas the ratio of the signal strength at 30 ft. to that at 10 ft. may be much less than 3:1.

621.396.81:621.397.5 3526
Field Test of U.H.F. Television—J. Fisher. (*Electronics*, vol. 22, pp. 106–111; September, 1949.) See also 1480 of June (Brown).

621.396.81:621.397.5 3527
Multipath Television Reflections—E. G. Hills. (PROC. I.R.E., vol. 37, pp. 1043–1046; September, 1949.) A formula is derived for the strength of reflected television signals at the receiver, and certain types of reflecting area are considered in order to discover why the direction of arrival of such signals depends on frequency. An antenna arrangement that appears to be particularly useful for reducing the effects of such reflections is described.

538.566+621.396.11 3528
Radio Wave Propagation [Book Review]—**Committee on Propagation of the National Defense Research Committee**—Publishers: Academic Press, New York, 1949, 511 pp. (PROC. I.R.E., vol. 37, p. 1036; September, 1949.) Effects at frequencies above 30 Mc are considered. Theoretical and practical aspects of standard propagation, nonstandard propagation, diffraction, coverage, radio-meteorology, absorption, scattering, and echoes are among the subjects covered. Wartime reports by 45 or more authors in various parts of the world are included.

RECEPTION

621.396.621 3529
F.M. Receivers with Supersonic Control—F. M. Berry. (*Communications*, vol. 29, pp. 12–14; August, 1949.) Control signals are used to enable a FM broadcaster to turn on or off specified groups of receivers, or to change the strength of signal transmitted to them. The control signals actuate locking circuits or switch out attenuator pads.

621.396.621 3530
Eddystone Model 680—(*Wireless World*, vol. 55, pp. 335–338; September, 1949.) Test report on a 15-tube superheterodyne general-purpose receiver for the frequency range 480 kc to 30 Mc.

621.396.621:338.585.8 3531
Reducing Costs in Receiver Manufacturing—S. A. Tucker. (*Electronics*, vol. 22, pp. 86–93; October, 1949.)

621.396.621:621.396.619.11/.13 3532
On Theoretical Signal-to-Noise Ratios in F.M. Receivers: A Comparison with Amplitude Modulation—D. Middleton. (*Jour. Appl. Phys.*, vol. 20, p. 724; July, 1949.) Corrections to 2619 of October.

621.396.621:621.396.619.11 3533
More on the Syndrome—D. G. Tucker and J. F. Ridgway. (*Short Wave Mag.*, vol. 5, pp. 598–601; December, 1947.) Abridged version of articles abstracted in 525 and 526 of 1948.

621.396.621:621.396.619.13:621.396.813 3534
Harmonic Distortion in Frequency-Modulation Off-Resonance Discriminator—A. R. Vallarino and M. S. Buyer. (*Elec. Commun.* (London), vol. 26, pp. 167–172; June, 1949.) The discriminator is assumed to consist of a single tuned circuit with a detector preceded by one or two limiters. The percentage nonlinearity and percentage distortion are shown graphically in terms of a parameter involving (a) the difference between the carrier frequency and the resonance frequency of the tuned circuit and (b) the width of the resonance curve at the 3-db level.

621.396.621.029.58 3535
Simultaneous Reception of [signals from] Several Stations on a Single Aerial—C. T. F. van der Wijck. (*Tijdschr. ned. Radiogenoot.*, vol. 8, pp. 365–393; March, 1941. In Dutch.) Description of suitable receiver circuits, their filters and transformers, with details of equipment for reception in the bands 13.5–20.5 Mc and 6.5–13.5 Mc.

621.396.621.54 3536
The Ultimate in Converters—J. E. Stacy. (*CQ*, vol. 5, pp. 13–20, 80; September, 1949.) A converter with remarkable performance at vhf is obtained by adding a triode mixer and a Clapp oscillator to the low-noise amplifier circuit described in 3061 of 1948 (Wallman, Macnee, and Gadsden).

621.396.822:523.72 3537
On the Origin of Solar Radio Noise—Haeff. (*See* 3412.)

621.396.828 3538
How VOA Combats Jamming—G. Q. Herrick. (*Electronics*, vol. 22, pp. 82–84; September, 1949.) The predistorter and peak limiter described accentuate speech intelligibility when multiple use of frequencies causes jamming.

STATIONS AND COMMUNICATION SYSTEMS

621.395.44 3539
Twelve-Channel Carrier Telephone Systems in South Africa—N. J. Paola, C. F. Boyce, and I. C. Ramsay. (*Trans. S. Afr. Inst. Elec. Eng.*, vol. 39, pp. 277–312; October, 1948. Discussion, pp. 313–316.) Crosstalk, noise, power supplies, and problems encountered in the planning, installation, and maintenance of these systems. See also 2610 of 1948 (Retief and Barker).

621.395.44:621.315.052.63 3540
Telecommunications at High Frequency—A. Latreille and M. Zweguintzow. (*Rev. Gén. Élec.*, vol. 58, pp. 349–356; September, 1949.) Discussion of systems suitable for use on hv power networks. Single-frequency, dual-frequency, and single-sideband equipment made by the Henry Lepaute Company and hf autocommutator equipment made by the Association des Ouvriers en Instruments de Précision are described and illustrated.

621.395.664.1 3541
Voice-Controlled Intercom System—J. R. Cooney. (*Electronics*, vol. 22, pp. 118, 140; September, 1949.) Operation is similar to that of the conventional master-substation system, except that when the operator at the master station speaks above a certain low threshold level, an amplifier is automatically switched on to transmit his words, and switched off when he stops speaking.

621.396.1:621.396.11.029.62 3542
The Ad Hoc Committee Report—Lewis. (*See* 3524.)

621.396.41 3543
Systems of U.S.W. Multichannel Telephony—W. Klein. (*Bull. Schweiz. Elektrotech. Ver.*, vol. 39, pp. 571–588; August 21, 1948. In German, with French summary.) The development of beam telephony in Switzerland and abroad is traced and the special characteristics of the different multichannel systems are reviewed, with particular reference to apparatus technique and to transmission properties. The choice of the best system for a specific service is still an open question. Points of particular interest, such as the optimum wave length range or the conditions required for linearity of a transmission channel, are studied in detail. Under certain practical assumptions, a formula is deduced for the mean power necessary per channel. Values of this mean power are calculated, as a function of the number of channels, for the different systems. The conditions in Switzerland are particularly suitable for the establishment of such systems between high-altitude stations. See also 546 of 1948 (Gerber and Tank).

621.396.41 3544
Some Development Work in connection with U.S.W. Multichannel Communications in Switzerland—G. Guanella. (*Bull. Schweiz. Elektrotech. Ver.*, vol. 39, pp. 678–688; October 2, 1948. In German, with French summary.) Describes experiments by Brown, Boveri and Co. on usw multicarrier systems, on systems using a single carrier and on modulated pulse systems.

621.396.41:621.392.52 3545
Significance and Application of Frequency Filters in U.S.W. Multichannel Systems—F. Staub. (*Bull. Schweiz. Elektrotech. Ver.*, vol. 39, pp. 627–635; September 18, 1948. In German with French summary.) By the use of frequency-selector filters, whatever the system of modulation, a very great number of channels can be obtained while keeping the expenditure of energy in each speech channel within reasonable limits, maintaining a favorable ratio between the transmitter energy and that radiated, and ensuring as far as possible reliability of operation. The design of suitable filters is discussed, their characteristics are described and actual filters, with their impedance transformers, are illustrated and performance figures given.

621.396.41:621.396.65 3546
A Time-Sharing System of Multiplex—H. D. B. Kirby. (*Electronic Eng.* (London), vol. 21, pp. 360–365; October, 1949.) About 4 samples per cycle must be transmitted for each channel, making about 250,000 samples per sec for a 25-channel system, including synchronizing pulses. Connection of each channel to the transmitting equipment is achieved by means of a number of on-off switches. Block diagrams of the transmitter and receiver are given, and waveform diagrams for various parts of the transmitter and receiver are discussed. Radio links of this type are operating satisfactorily in Holland, South Africa, and Czechoslovakia.

621.396.619 3547
Comparison of Signal/Noise Ratios of Modulation Arrangements—W. Runge. (*Arch. Elek.* (Übertragung), vol. 3, pp. 155–159; August, 1949.) A method of expressing signal-to-noise ratios in terms of two parameters is explained and applied to various modulation systems. The ratios are tabulated for single-sideband systems with carrier suppression, double-sideband systems with carrier, FM, and pulse-phase, pulse-amplitude, pulse-time, and pulse-code modulation. See also 3552 below.

621.396.619.13 3548
Application of Negative Feedback to Frequency-Modulation Systems—P. F. Panter and W. Dite. (*Elec. Commun.* (London), vol. 26, pp. 173–178; June, 1949.) A general formula for the output voltage in terms of the applied voltage is derived for a nonlinear network with feedback. For a FM transmitter, feedback will reduce distortion provided that a limiter is used in the feedback path. For a receiver, a limiter is required before the discriminator to obtain constant output. Inverse feedback reduces the required bandwidth of the if amplifier without sacrificing the advantage of wide-band FM.

621.396.619.13 3549
Recent Developments concerning Frequency Modulation—T. J. Weijers. (*Tijschr. ned. Radiogenoot.*, vol. 8, pp. 315–364; October, 1940. In Dutch.) A comprehensive discussion dealing with the response of a linear network to a FM signal, various methods of detection, the balanced detector, amplitude limiting and the use of feedback in FM receivers, effects of interference and noise, and interference from a neighboring transmitter on the same frequency. The relative merits of AM and FM for medium and for short waves are summarized.

621.396.619.16 3550
Note on the Theoretical Efficiency of Information Reception with P.P.M. [pulse position modulation]—M. J. E. Golay. (Proc. I.R.E., vol. 37, p. 1031; September, 1949.)

621.396.619.16 3551
Equipment for Interference-Free Long-Distance Electrical Communications—F. Schröter. (*Arch. Elek.* (Übertragung), vol. 1, pp. 2–13; July and August, 1947.) Description of a pulse-code system in which signal amplitudes are converted into pulse groups in the transmitter by means of an electro-optical system including a cr tube and a screen with graded sets of perforations. In the receiver the process is reversed, the deflection of the beam in the cr tube being synchronized with that in the transmitter.

621.396.619.16 3552
Noise Factor for Communication by Code Modulation—E. Kettel. (*Arch. Elek.* (Übertragung), vol. 3, pp. 161–164; August, 1949.) The effect of noise in producing signal imperfections in pulse-code modulation systems is discussed. Signal-to-noise ratios for pulse-code, pulse-phase, and frequency modulation are calculated from Runge's formulas (3547 above) and compared.

621.396.932 3553
R.M.S. Caronia Radio and Electronic Installation—(*Elec. Commun.* (London), vol. 26, pp. 107–128; June, 1949.) Detailed account of the equipment briefly noted in 1799 of July.

621.394/.395:621.316.975 3554
Earth Conduction Effects in Transmission Systems [Book Review]—E. D. Sunde. Publishers: D. Van Nostrand, New York, 1949, 360 pp., $6.00. (Proc. I.R.E., vol. 37, p. 1034; September, 1949.) Fundamental methods for the analysis of such effects and the basic principles underlying devices for protection against the resultant circuit disturbances are discussed. "... This book is the first of its kind, covering a difficult field very well and filling a need long neglected."

SUBSIDIARY APPARATUS

621.316.722.1 3555
Voltage Stabilizers: Parts 1–4—F. A. Benson. (*Electronic Eng.* (London), vol. 21, pp. 155-158, 200–203, 243–247, and 300–302; May and August, 1949.) A review of various methods of stabilization for both ac and dc power supplies. Circuits including magnetically-saturated elements are considered as well as those using gas-

eous discharge tubes. Characteristics of available tubes and their limitations are discussed. Several tube circuits found useful in practice are described, with due reference to the performance to be expected. The treatment is nonmathematical, except for an appendix on tube-circuit theory.

621.396.682 3556
A D.C. Stabilized Power Supply of Low Impedance—V. H. Attree. (*Jour. Sci. Instr.*, vol. 25, pp. 263–268; August, 1948.) A full-wave rectifier with a single-stage π filter followed by a series power tube. Output impedance is $<2\Omega$ at all frequencies up to 100 kc or more. For load currents of 0–90 mA the output is 400 V, with variations of 200 mV.

TELEVISION AND PHOTOTELEGRAPHY

621.397.26 3557
New Methods of Rapid Phototelegraphy—F. Schröter. (*Bull. Schweiz. Elektrotech. Ver.*, vol. 39, pp. 819–827; December 11, 1948. In German, with French summary.) Cathode-ray tubes are used for scanning and reproducing the pictures to be transmitted. The screen of the transmitter tube has very little persistence, while the receiving screen is highly persistent. Fidelity and image sharpness are secured by correction of distortion, automatic compensation of the field curvature and automatic focusing. Using a 1200-line sweep, a picture 20 cm $\times$14 cm can be transmitted on any type of wide band system in about 1 sec.

621.397.26:621.394 3558
Facsimile Transceiver for Pickup and Delivery of Telegrams—G. H. Ridings. (*Elec. Commun.* (London), vol. 26, pp. 129–137; June, 1949.) Description of the Western Union Telefax system.

621.397.331.2:535.517.25 3559
Asymmetrical-Aperture Scanning in Television—Ya. A. Ryftin. (*Zh. Tekh. Fiz.*, vol. 19, pp. 804–821; July, 1949. In Russian.) A new scanning method which increases image sharpness by about 85 per cent and more than doubles the "quality of transmission." No increase of bandwidth is required, but the resolving power of the camera must be increased. See also 1785 of 1948.

621.397.5:061.3 3560
The Paris Television Congress—(*Jour. Telev. Soc.*, vol. 5, pp. 238–244; December, 1948.) Summaries of papers noted in 865 of April.

621.397.5:535.88:532.62 3561
Large-Screen Television and the Eidophor Process—H. Thiemann. (*Télév. Franç.*, no. 50, pp. 6–10, 37; August, 1949.) See also 296 of 1948.

621.397.5:535.88:621.385.832 3562
On the Suitability of the Cathode-Ray Tube with Fluorescent Screen for Television Projection in Cine-Theatres—F. Fischer. (*Bull. Schweiz. Elektrotech. Ver.*, vol. 39, pp. 468–480; July 24, 1948. In German, with French summary.) The luminous intensity required for large-screen projection is discussed. The luminous efficiency of fluorescent substances at present available has almost reached the theoretical upper limit. The construction of fluorescent screens to furnish maximum light for a given definition of the image, and possible means of developing still greater luminous intensities, are considered.

621.397.5:621.396.81 3563
Field Test of U.H.F. Television— Fisher. (*See* 3526.)

621.397.5:621.396.81 3564
Multipath Television Reflections—Hills. (*See* 3527.)

621.397.5:621.396.81 3565
Technique for TV Field Surveys—Dreyer. (*See* 3525.)

621.397.5(083.74) 3566
Present-Day Television Standards in the U.S.A.—W. Reichel. (*Arch. Elek.* (Übertragung), vol. 3, pp. 175–181; August, 1949.)

621.397.5(083.74) 3567
Progress toward International TV Standards—D. G. Fink. (*Electronics*, vol. 22, pp. 69–71; October, 1949.) At the first meeting of the 11th Study Group of the C.C.I.R., substantially unanimous agreement was reached that (a) the aspect ratio should be 4 units horizontally to 3 units vertically, (b) interlacing should be adopted, using the two-to-one odd-line method, and (c) vertical scanning should be independent of the frequency of the power system. Further study is needed for determining suitable regional or world-wide standards for the number of lines per frame, number of frames per second, channel width and related quantities; existing national standards for these quantities vary considerably.

621.397.62 3568
Converters for U.H.F. Television Reception—D. K. Reynolds and M. B. Adams. (*Electronics*, vol. 22, pp. 92–96; September, 1949.) Designed to cover the 475–890-Mc band with an if of 250 Mc, the local oscillator tuning range being 271–680 Mc. This tuning range is obtained with a circuit including a semibutterfly oscillator, a tap-switch oscillator, and a cylinder oscillator. A crystal mixer is used because of its low noise figure at the required frequencies. This mixer consists essentially of a rolled-up parallel-strip transmission line. Parallel-line and coaxial-cavity types of mixer are suitable for fixed-tuned converters.

621.397.62:535.88:621.397.335 3569
Projection-Television Receiver: Part 5—The Synchronization—J. Haantjes and F. Kerkhof. (*Philips Tech. Rev.*, vol. 10, pp. 364–370; June, 1949.) Part 4: 2934 of November.

621.397.621 3570
Anastigmatic Yoke for Picture Tubes—K. Schlesinger. (*Electronics*, vol. 22, pp. 102–107; October, 1949.) 1949 IRE Convention paper noted in 1810 of July [No. 85]. Loss of picture detail at the corners of large cr tubes is caused primarily by spot or deflection defocusing. It can be corrected by means of a deflection yoke that provides deflection fields which increase with distance from the tube axis. Such a yoke introduces only a tolerable amount of barrel distortion. Correction is easier for tubes with magnetic deflection than for those with es deflection. A two-dimensional theory of deflection-coil astigmatism is explained and practical correction arrangements based on this theory are described.

621.397.7 3571
WOR-TV F.M.—F. J. Bingley. (*Electronics*, vol. 22, pp. 70–81; September, 1949.) Description of the television transmitter, associated terminal equipment to enable television pictures to be brought to the transmitter, and of an associated FM broadcast transmitter. An 810-ft tower supports both the television and FM antennas.

621.397.7:621.317.79 3572
A TV Monitor Receiver—Grace. (*See* 3488.)

621.397.7.001.4:621.396.81 3573
TV Site Testing and Measurement Techniques—E. S. Clammer. (*Communications*, vol. 29, pp. 6–9; June, 1949.) A pulse transmitter whose power output is about equal to that of the proposed television station is used. Field intensities and the position of objects producing echoes are observed. A helicopter provides the best means of supporting the test antenna and transmitter. The transmitting and mobile receiving equipment used is described and typical results are discussed; adequate qualitative data can be obtained to determine correctly whether good, mediocre, or unsatisfactory service would be given.

621.397.82 3574
The Character of Interference Patterns in Television—G. G. Gouriet. (*Jour. Telev. Soc.*, vol. 5, pp. 235–237; December, 1948.) Diagonal patterns are caused by the presence of an interfering carrier signal whose frequency is within the vision frequency band. The change in these patterns with variation of the interfering frequency is analyzed.

621.383+621.397 3575
Photoelectricity and its Applications [Book Review]—Zworykin and Ramberg. (*See* 3599.)

TRANSMISSION

621.396.61:621.316.726 3576
Instantaneous Deviation Control—M. R. Winkler. (*Electronics*, vol. 22, pp. 97–99; September, 1949.) The audio signal of a FM transmitter is differentiated, amplified until at a level suitable for clipping, clipped by a pair of biased diodes, and integrated so that the output wave is identical with the input wave except for slope limiting. Transmitter frequency deviations are held within definite limits determined by the maximum allowable slope.

TUBES AND THERMIONICS

537.533 3577
Electron Beams of High Current-Density in Electrostatic Fields—H. Huber and W. Kleen. (*Arch. Elektrotech.*, vol. 39, pp. 394–414; April, 1949.) Detailed theory and discussion with special reference to the design of electron guns.

621.383.4 3578
Theory of Photoconductivity of Layers of Semiconducting Substances—E. Schwarz. (*Proc. Phys. Soc.*, vol. 62, pp. 530–532; August 1, 1949.) The adsorption of oxygen ions during the evaporation of semiconducting substances and after the deposit has been formed is an essential condition for the production of photoconductive layers. A qualitative theory based on experimental results is proposed. See also 1102 of May.

621.385.001.4 3579
Modern Test Methods revealing the Characteristics and Limitations of V.H.F. and U.H.F. Valves—Remillon. (*See* 3489.)

621.385.029.63/.64+621.396.615.141.2 3580
On the Interaction between an Electron Beam and Travelling Electromagnetic Waves—W. Kleen. (*Elektron Wiss. Tech.*, vol. 3, pp. 341–349; September, 1949.) A concise account of the theory, construction, and properties of the traveling-wave tube. Interaction in the magnetron is also considered.

621.385.029.63/.64 3581
Investigations of Self-Excited Oscillations in Travelling-Wave Valves—H. Schnitger and D. Weber. (*Frequenz*, vol. 3, pp. 189–195; July, 1949.) Well-defined powerful oscillations of wavelength 5 to 50 cm were obtained in ordinary traveling-wave tubes without any external circuit connections. Oscillation regions were established whose mean wavelength increased with the applied voltage. Within an oscillation region the wavelength falls slightly with decreased voltage; for $\lambda=24.6$ cm the frequency variation is about 50 kc per volt. The results are discussed with reference to Pierce's theory (2284 of 1947). Research with tubes having the spiral outside the glass wall of the tubular portion shows interesting possibilities.

621.385.029.63/.64:621.396.645 3582
The Double-Stream Amplifier—A. V. Hollenberg. (*Bell. Lab. Rec.*, vol. 27, pp. 290–292; August, 1949.) See 2487 of October.

621.385.029.63/.64:621.396.645 3583
Double-Stream Amplifiers—Pierce. (*See* 3387.)

621.385.029.63/.64:621.396.65 3584
Travelling-Wave Amplifier for 6 to 8 Centimetres—D. C. Rogers. (*Elec. Commun.* (London), vol. 26, pp. 144–152; June, 1949.) A review of development work which demonstrated the usefulness of the traveling-wave tube as an unattended repeater-amplifier in a microwave communication link. Tubes with higher output for the power stages and with lower noise factor for the input stages still need to be developed.

621.385.032.21 3585
Thermionic Emission from Sintered Cathode of Thoria and Tungsten Mixture—H. Y. Fan. (*Jour. Appl. Phys.*, vol. 20, pp. 682–690; July, 1949.) An experimental investigation for a cathode sintered from a mixture of 67 per cent thoria and 33 per cent tungsten. Emission was found to be somewhat lower than that of cathodes sintered from pure thoria. Change of cathode activity with temperature was studied in detail, and a satisfactory theoretical explanation is given.

621.385.032.213 3586
Potassium-Activated Cold-Cathode Tubes—A. L. Chilcot and F. G. Heymann. (*Jour. Sci. Instr.*, vol. 26, pp. 289–294; September, 1949.) For most applications of cold-cathode tubes, uniformity of characteristics between large numbers of tubes of the same design, and for individual tubes over long periods of time, is important. This uniformity depends upon accurate reproduction of (a) filling-gas pressure, (b) electrode spacings, and (c) the cathode surface. The results obtained for a diode with nickel electrodes, the cathode being coated with potassium, are shown graphically; the curve for clean potassium is accurately reproducible; the contaminating effect of oxygen and water vapor is shown. The leads for the cathode and for the other electrodes of these tubes are brought out at opposite ends of the glass envelope; otherwise a conducting potassium film between different electrodes is difficult to avoid. Design details for various types of tubes for given applications are discussed.

621.385.032.216 3587
Oxide Cathode Theory—W. Couch. (*Electronics*, vol. 22, pp. 190, 198; October, 1949.) Brief discussion of the mechanism of activation. See also 1028 of 1940 (Blewett).

621.385.032.216 3588
Reversibility of Oxygen Poisoning in Oxide-Cathode Valves—G. H. Metson. (*Nature* (London), vol. 164, pp. 540–541; September 24, 1949.) The total emission recovered its initial value after three successive poisonings by oxygen sufficient to reduce emission by over 90 per cent. When a diode was arranged so that the polarity of the anode voltage could be reversed instantaneously, recovery was found to be much more rapid while current was passing through the cathode than when current was not passing. A cathode is more resistant to poisoning at higher current densities. Gas pressure inside the tube increases during recovery.

621.385.032.216 3589
Note on Volt-Dependent Poisoning Effects in Oxide-Cathode Valves—G. H. Metson. (*Proc. Phys. Soc.*, vol. 62, pp. 589–591; September 1, 1949.) Cathode poisoning can be due to bombardment by electrons with certain discrete energies. Possible explanations of such poisoning at electron energies of 5.56, 10 and 15.9 eV are given.

621.385.032.216 3590
D.C. and Pulsed Emission from Oxide Cathodes—D. A. Wright. (*Proc. Phys. Soc.*, vol. 62, pp. 398–400; June 1, 1949.) Continuation of 2091 of August. Further investigations are reported on the drop in emission with time under continuous dc operation. The effect is shown to be due to gas poisoning by ions returning from the anode or the anode/cathode space to the cathode. The results support the view that pulsed emission is derived by thermal ionization of adsorbed Ba as in de Boer's theory.

621.385.032.216 3591
On Barium-Sulphide Layers on Oxide Cathodes and their Influence on the Emission—H. A. Stahl. (*Schweiz. Arch. Angew. Wiss. Tech.*, vol. 14, pp. 337–343; November, 1948.) Electron-diffraction diagrams reveal the presence of BaS in oxide-cathode layers prepared from $BaCO_3$ or from carbonate mixtures. The sulphur is probably derived from sulphurous constituents of the atmosphere. Emission measurements show that the presence of BaS has a deleterious effect. It appears possible that differences in the performance of individual tubes of the same type may be due to the existence of BaS in differing amounts in the coatings of their cathodes.

621.385.032.216 3592
Use of Radioactive Elements for investigating the Behaviour of the Alkaline-Earth Metals of Oxide Cathodes—J. Beydon, L. Beaudoin, J. Challansonnet, and J. Debiesse. (*Compt. Rend. Acad. Sci.* (Paris), vol. 229, pp. 353–354; August 1, 1949.) The volatilization of the oxides covering the cathodes was studied by incorporating in these oxides radioactive Ba^+ and Sr^+, and measuring with a G-M counter the activity obtained on the different electrodes of the EL3 tubes used, at various stages of their operation. The loss of Ba and Sr from the cathodes occurs largely during their formation and activation.

621.385.032.216 3593
Influence of the Density of Emission on the Life of Oxide Cqthodes—S. Wagener. (*Nature* (London), vol. 164, pp. 357–358; August 27, 1949.) Cathode life is not prejudiced by a considerable increase of emission density, provided ionization within the tube is prevented. Experimental results are discussed.

621.385.032.216:621.315.59 3594
Semi-Conducting Properties in Oxide Cathodes—Hannay, MacNair, and White. (*See* 3444.)

621.385.032.216:621.317.79 3595
A New Emission Microscope for Oxide Cathodes—Jacob. (*See* 3486.)

621.396.615.14 3596
On the Generation of Electromagnetic Oscillations in a Spiral by an Axial Electron Current—B. B. van Iperen. (*Philips Res. Rep.*, vol. 4, pp. 20–30; February, 1949.) The exchange of energy between an electron and an alternating electric field is analyzed mathematically. The theory is applied to the interaction of a current in a spiral with an electron stream flowing along the axis, conditions for oscillation being deduced. Experimental results are in agreement with the theory. See also 893 of 1948 (Fremlin *et al.*).

621.396.621.53:537.311.33:621.315.59 3597
A Crystal Tetrode Mixer—R. W. Haegele. (*Sylvania Technologist*, vol. 2, pp. 2–4; July, 1949.) Construction details are given. The conversion transconductance is equal to that of mixer tubes and operation is satisfactory at frequencies up to 200 Mc. The isolation of the input circuit is better than with diode or triode mixers. See also 3598 below.

621.396.621.53:537.311.33:621.315.59 3598
Crystal-Tetrode Mixer—R. W. Haegele. (*Electronics*, vol. 22, pp. 80–81; October, 1949.) For another account see 3597 above.

621.396.645:537.311.33:621.315.59 3599
On the Hole Current in the Germanium Transistor—K. Lehovec. (*Phys. Rev.*, vol. 75, p. 1100; April 1, 1949.)

621.383+621.397 3600
Photoelectricity and its Applications [Book Review]—V. K. Zworykin and E. G. Ramberg. Publishers: J. Wiley and Sons, New York, 1949, 494 pp., $7.50. (*Electronics*, vol. 22, pp. 250, 252; October, 1949.) "This book covers virtually the entire field of photoelectricity from its historic beginning to its recent application in Ultrafax."

621.385.832 3601
Cathode Ray Tube Displays [Book Review]—T. Soller, M. A. Starr, and G. E. Valley, Jr. (Eds.). Publishers: McGraw-Hill, New York and London, 1948, 746 pp., 60 s. (*Nature* (London), vol. 164, pp. 594–595; October 8, 1949.) Vol. 22 of the M.I.T. Radiation Laboratory series. A very comprehensive treatment of the principles of operation of the cr tube and its use for displaying various types of radar and other information. A chapter on screen materials is included.

621.396.615.142 3602
Modern Radio Technique]Book Review]—A. H. W. Beck. Publishers: Macmillan, New York, 1948, 173 pp. (PROC. I.R.E., vol. 37, p. 1038; September, 1949.) The operation of vm tubes is described with considerable attention to design.

MISCELLANEOUS

621.39:061.3 3603
The International Conferences of the Union Internationale des Télécommunications—A. Möckli. (*Bull. Schweiz. Elektrotech. Ver.*, vol. 39, pp. 607–612; September 4, 1948. In French, with German summary.) The objects and the organization of the Union are explained; a brief review of the work of individual conferences and committees shows the immense progress in international collaboration realized in recent years. The great technical and linguistic difficulties which complicate the work of international conferences are discussed and the way in which these difficulties are overcome by good organization and technical measures is outlined. A short report of the results of the conferences at Atlantic City is included.

621.396 3604
Outline of Radio [Book Review]—Publishers: G. Newnes, London, 21s. (*Engineering* (London), vol. 168, p. 144; August 5, 1949.) "This volume, of nearly 700 pages, has been compiled by a team of eight contributors, and describes in simple terms the electrical phenomena and principles upon which are founded the allied techniques of radio, television, and radar."

ABSTRACTS AND REFERENCES INDEX

The Index to the Abstracts and References published throughout the year is in course of preparation, and will, it is hoped, be available in February from *Wireless Engineer*, price 2s. 8d. (including postage). As supplies are limited the Publishers ask us to stress the need for early application for copies. Included with the Index is a selected list of journals scanned for abstracting, with publishers' addresses.